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ПО ПАТЕНТАМ И ТОВАРНЫМ ЗНАКАМ

(12) ОПИСАНИЕ ИЗОБРЕТЕНИЯ К ПАТЕНТУ РОССИЙСКОЙ ФЕДЕРАЦИИ

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(98) Адрес для переписки:
129010, Москва, ул. Большая Спасская, 25, стр.3, ООО "Юридическая фирма Городисский и Партнеры", пат.пов. Ю.Д.Кузнецову, рег.№ 595

(71) Заявитель:
ОЛИВЕТТИ ТЕЛЕМЕДИА С.П.А. (IT)

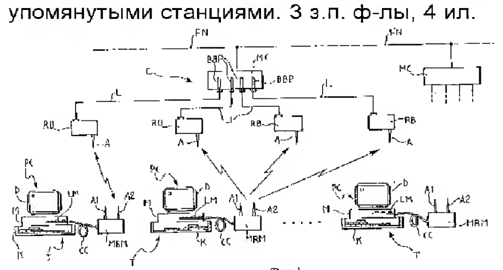
(72) Изобретатель: **БАД Эндрю (GB), ФУРНО Рино (IT)**

(73) Патентообладатель:
ОЛИВЕТТИ ТЕЛЕМЕДИА С.П.А. (IT)

(74) Патентный поверенный:
Кузнецов Юрий Дмитриевич

(54) СПОСОБ ОБМЕНА ДАННЫМИ МЕЖДУ МНОЖЕСТВОМ АБОНЕНТСКИХ СТАНЦИЙ ПО БЕСКАБЕЛЬНОЙ ЛОКАЛЬНОЙ СЕТИ ЧЕРЕЗ ЦЕНТРАЛЬНУЮ УПРАВЛЯЮЩУЮ СТАНЦИЮ

(57)
Изобретение относится к способу передачи данных по радио в соответствии со стандартом DECT, использующим широкую полосу частот, разделенную на множество каналов и на заданное число временных сегментов. Техническим результатом является разработка способа, позволяющего усовершенствовать работу известных бескабельных сетей. Технический результат достигается тем, что из центральной станции периодически транслируют сигналы для каждого канала, а абонентские станции периодически сканируют эти сигналы и определяют уровни сигналов каждого из множества каналов для выявления занятых или незанятых временных сегментов. Далее с помощью каждой из абонентских станций формируют и периодически корректируют список значений уровней сигналов и незанятых временных сегментов, после чего устанавливают радиосвязь между выбранной абонентской станцией и центральной управляющей станцией и при наличии информации, подлежащей передаче, производят обмен информацией между упомянутыми станциями. 3 з.п. ф-лы, 4 ил.



Фиг 1

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(98) Mail address:
129010, Moskva, ul. Bol'shaja Spasskaja, 25, str.3, OOO "Juridicheskaja firma Gorodisskij i Partnery", pat.pov. Ju.D.Kuznetsovu, reg.№ 595

(71) Applicant:
OLIVETTI TELEMEDIA S.P.A. (IT)

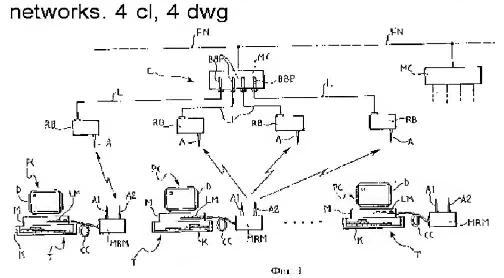
(72) Inventor: BAD Ehndrju (GB), FURNO Rino (IT)

(73) Proprietor:
OLIVETTI TELEMEDIA S.P.A. (IT)

(74) Representative:
Kuznetsov Jurij Dmitrievich

(54) METHOD FOR DATA EXCHANGE AMONG PLURALITY OF SUBSCRIBER STATIONS OVER WIRELESS LOCAL NETWORK THROUGH CENTRAL CONTROL STATION

(57) Abstract:
FIELD: data transmission over radio link. SUBSTANCE: method is used for data transmission according to DECT standard using broad frequency band divided into plurality of channels and desired number of time segments. Signals are periodically sent from central station to each channel and subscriber substations periodically scan these signals and determine signal levels in each of plurality of channels to detect busy or vacant time segments. Then list of signal levels and vacant time segments is compiled and periodically corrected by means of each subscriber station whereupon radio communication is established between chosen subscriber station and central control station; if data to be transmitted is present, data exchange between mentioned stations takes place. EFFECT: enhanced operating reliability of modified wireless networks. 4 cl, 4 dwg



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Изобретение относится к локальной сети (LAN), а более конкретно к способу обмена данными между множеством абонентских станций по бескабельной сети посредством неподвижного центрального управляющего устройства, причем каждая из абонентских станций содержит терминал для ввода данных.

Локальные сети получили широкое распространение в сфере информатики и видеографической связи при установлении связи на небольших расстояниях с целью обеспечения передачи и распределения данных и услуг между множеством пользователей, находящихся на одном и том же участке, например, в одном здании. Локальная сеть дает возможность использовать множество разного рода терминалов ввода данных, таких, как персональные компьютеры (PC), миникомпьютеры, принтеры и так далее, которые могут присоединяться чрезвычайно гибким образом, обеспечивая повышенную скорость передачи порядка сотен тысяч килобит в секунду.

До настоящего времени в основном использовались локальные сети беспроводного типа, то есть сети, в которых соединения между станциями пользователей и центральными управляющими устройствами целиком реализуются с помощью проводов.

Появление на рынке портативных компьютеров, таких, как портативные персональные компьютеры, определило потребность в бескабельных локальных сетях.

Бескабельная локальная сеть уменьшает затраты на установку, поскольку исключает необходимость в установке соединительных кабелей. Сеть подобного типа может также формироваться в случаях, когда трудно или невозможно установить соединительные провода, например, в условиях отсутствия розеток для локальных сетей, либо при наличии архитектурных ограничений.

Бескабельная локальная сеть может представлять идеальное решение в организации, в которой расположения станций пользователей или число станций, связанных в сеть, подвергаются частым изменениям или модификациям.

Бескабельная локальная сеть представляет также идеальное решение для организаций, которые часто меняют свое местоположение. В этом случае было бы фактически непрактично и неэкономично переносить проводную локальную сеть.

Наконец, как говорилось выше, локальная сеть обеспечивает возможность обмена данными даже портативных персональных компьютеров, без ограничения подвижности этих новых устройств.

Соответствующая изобретению сеть работает, в частности, согласно стандарту DECT (Digital European Cordless Telecommunications - Цифровые европейские бескабельные телекоммуникации), разработанному ETSI, Европейским институтом телекоммуникационных стандартов, который определяет спецификации для радиосвязи между пользователями и сетью в условиях частного окружения.

Система DECT работает в полосе частот 1880-1900 МГц и обеспечивает радиопередачу посредством гибридной системы с временным и частотным уплотнением каналов.

Характеристики стандарта DECT описываются, например, в Digital European Cordless Telecommunications Services and Facilties (Цифровые европейские телекоммуникационные услуги и средства), ETSI DR/RES 3003, за июнь 1991 года и в "Data Services in DECT", A. Bud (Обслуживание данных в DECT, Э. Бад), Пятая международная конференция Института инженеров по электронике по наземной передвижной радиосвязи, Уорик, декабрь 1989 года.

Беспроводная локальная сеть, использующая радио для установления связи между множеством станций пользователей, где каждая содержит соответствующий терминал ввода данных, посредством неподвижного центрального управляющего устройства, которое управляет связью между терминалами ввода данных в соответствии с предварительно определенным стандартом связи, раскрывается в EP-A-0257947. В этой известной бескабельной локальной сети каждый терминал ввода данных связывается с отдельным неподвижным радио-приемопередатчиком, а центральное управляющее устройство соединяется с неподвижными радиобазами.

Сетевые системы для передачи данных радиосигналами между основным приемным устройством и множеством рабочих станций раскрываются в PATENT ABSTRACTS OF JAPAN (Японские патентные аннотации), том 14, номер 229 (E-928), 4172, 15 мая 1990 г. и JP-A-0260252.

В основу изобретения поставлена задача разработать способ, позволяющий усовершенствовать работу известных бескабельных локальных сетей.

Поставленная задача решается тем, что в способе обмена данными между множеством абонентских станций по бескабельной локальной сети через центральную управляющую станцию, в соответствии с которым широкую полосу частот делят на множество (n) каналов (f_1-f_{10}) и на заданное число временных сегментов (2m), согласно изобретению периодически транслируют с центральной управляющей станции (C) сигналы для каждого из множества (n) каналов (f_1-f_{10}) и для каждого из временных сегментов (2m), периодически сканируют множеством абонентских станций (T) указанные сигналы с целью определения уровня сигналов каждого из множества (n) каналов (f_1-f_{10}) и факта занятости, либо незанятости временных сегментов (2m), формируют и периодически корректируют с помощью каждой из абонентских станций (T) список значений уровней сигналов и незанятых временных сегментов, устанавливают радиосвязь между одной из выбранных абонентских станций (T) и центральной управляющей станцией (C) в случае, когда выбранная абонентская станция (T) содержит информацию, подлежащую передаче, производят обмен информацией между указанной выбранной абонентской станцией (T) и центральной управляющей станцией (C) путем выбора из числа незанятых временных сегментов с уровнем сигнала, обеспечивающим оптимальное

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отношение сигнал/шум.

Целесообразно поддерживать радиосвязь в течение заданного периода времени до завершения обмена информацией.

Целесообразно также поддерживать радиосвязь в течение адаптивно заданного периода времени на основе статистических данных связной нагрузки относительно выбранной абонентской станции (T), полученных в заранее определенный период времени.

Предпочтительно обеспечивать асимметричную многоканальную широкополосную связь с использованием множества сегментов из заданного числа временных сегментов (2m), используемых одновременно при установлении указанной радиосвязи.

Обычно терминалами ввода данных у станций пользователей могут быть, например, персональные компьютеры, а макропро-цессорное адаптерное устройство для удобства изготавливается в виде имеющей формат "половинного размера" карты или дочерней платы, встроенной в персональный компьютер и связанной с его шиной. Адаптер, таким образом, запитывается от шины терминала ввода данных, что обеспечивает дополнительные удобства.

Кроме того, радиомодуль передатчика и приемника запитывается от связанной с ним платы адаптера посредством проводников, которые тянутся через гибкий многожильный кабель, соединяющий его с платой, что является дополнительным преимуществом.

Еще одно преимущество заключается в том, что радиомодуль передатчика и приемника каждой абонентской станции имеет две ненаправленные антенны для получения пространственного "разноса" для улучшения характеристик радиосвязи с неподвижными радиомодулями или базами.

Может также монтироваться неподвижное центральное управляющее устройство для связи с неподвижной сетью, например, с сетью Ethernet, с кольцевой сетью с эстафетным доступом или с сетью через RS232.

В дальнейшем изобретение поясняется описанием варианта его осуществления со ссылками на прилагаемые чертежи, в числе которых:

фиг.1 изображает блок-схему локальной сети,

фиг. 2 - блок-схему, показывающую структуру адаптера и подвижного радиомодуля, связанного с каждым терминалом ввода данных локальной сети, показанной на фиг. 1,

фиг. 3 - частотно-временную диаграмму, описывающую процесс радиопередачи согласно гибридной системе с временным и частотным уплотнением каналов в локальной сети по фиг. 1,

фиг.4 - пример кадра для асимметричного соединения с множеством однонаправленных каналов, которое может образоваться в локальной сети по фиг. 1.

На фиг.1 бескабельная локальная сеть LAN, образованная в соответствии со спецификациями стандарта DECT, включает в себя множество абонентских станций T и неподвижное центральное управляющее устройство, обычно обозначаемое C.

Каждая абонентская станция T содержит соответствующий терминал ввода данных, который в общем случае может состоять из любого устройства, такого, как процессор, принтер и так далее, которое может посылать и/или принимать цифровые данные посредством сети связи. В примере согласно фиг.1 терминалы ввода данных абонентских станций T образованы персональными компьютерами PC, имеющими стандартную сеть и программное обеспечение типа LAN Manager. Персональными компьютерами могут быть, например, устройство Olivetti I/D33, где каждое включает клавиатуру K, экран дисплея D в модуль обработки данных M.

Каждый терминал ввода данных соединяется с соответствующим подвижным радиомодулем передатчика и приемника (приемопередатчика), обозначенным MRM, типа, отвечающего спецификациям DECT для физического уровня.

Модуль обработки данных M каждого терминала ввода данных PC содержит в себе соответствующее микропроцессорное устройство адаптера, обозначаемое LM. Микропроцессорный адаптер адаптирован к режиму работы в качестве интерфейса между соответствующим терминалом ввода данных и связанным с ним подвижным радиомодудем MRM. С этой целью, как схематически показано на фиг. 2, микропроцессорный адаптер LM соединяется с шиной данных DB модуля обработки данных M терминала ввода данных. Адаптер LM соединяется также с подвижным радиомодулем MRM, связанным с терминалом ввода данных посредством многожильного гибкого кабеля CC (фиг. 1 и 2).

Центральное управляющее устройство C включает в себя множество неподвижных радиомодулей или баз FRM, установленных в соответствующих предварительно определенных фиксированных местах для передачи и приема пакетов данных подвижным радиомодулем MRM одной или нескольких абонентских станций T и от него.

Радиобазы RB присоединяются, например, электрическими проводами L к микропроцессорному концентратору MC, который устанавливается в фиксированном месте и программируется на управление связью между абонентскими станциями T по предварительно определенным процедурам и протоколам, в соответствия со стандартом DECT, посредством радиосвязи, установленной между подвижными радиомодулями MRM и радиобазами RB.

Предпочтительно, чтобы мог монтироваться концентратор MC для связи с неподвижной сетью FN, например, с сетью Ethernet, с кольцевой сетью с эстафетным доступом или с сетью через RS232. Может оказаться возможным присоединение к неподвижной сети концентраторов MC других локальных сетей LAN.

Интегрированная система, описанная со ссылкой на фиг. 1, может выполнять функцию многопортового моста уровня управления доступом к среде (MAC) для обеспечения передачи и приема абонентскими станциями T пакетов данных, которые упаковываются в соответствии с форматом стандарта DECT и подвергаются обмену по радио посредством неподвижной части C системы. Эта часть действует в качестве высокоскоростной

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системы коммутации пакетов и направляет принимаемые пакеты пользовательским станциям места назначения или проводной сети FN.

Описываемая система работает в соответствии со стандартом DECT. Отвечающее стандарту DECT соединение между абонентскими станциями T и неподвижной частью C системы заменяет только уровень управления доступом к среде (MAC) системы Ethernet.

Благодаря линиям L радиобазы RB могут устанавливаться от концентратора MC на расстояниях до порядка 100 м. Путем выполнения функций, таких, как передача обслуживания соединений (handover), которые предусмотрены для стандарта DECT, может устанавливаться почти полная непрерывность обслуживания между двумя или несколькими используемыми радиобазами RB.

Концентратор MC может быть сформирован, например, на основе персонального компьютера Olivetti M300 с процессором Intel 386Sx, работающим с тактовой частотой 16 МГц.

Этот концентратор включает в себя процессоры групповых сигналов BBP, упорядоченным образом соединенные с соответствующими связанными радиобазами RB.

Удобно, что процессоры групповых сигналов BBP концентратора MC и интерфейсные адаптеры LM абонентской станции T могут быть выполнены в виде монтажных плат персонального компьютера с форматом половинного размера и на практике могут иметь ту же самую структуру на аппаратурном уровне и отличаться только на уровне программного обеспечения. Структура интерфейсного адаптера LM станции пользователя более подробно будет описана ниже со ссылкой на фиг. 2.

Концентратор MC в целом обеспечивает управление всей системой и, в частности:

- функционирование высоких уровней протоколов DECT,
- управление различными ресурсами сети,
- переключение пакетов данных и, в случае необходимости,
- сопряжение между бескабельной сетью LAN и проводной сетью FN.

Высокие уровни протоколов DECT обеспечивают услуги, такие, как высокоскоростная передача обслуживания, опознавание пользователя и создание виртуальных соединений, которые обеспечивают установление физических соединений без массивных обменов данными.

Перед дальнейшим обсуждением достоинств структуры функций устройств LM и процессоров групповых сигналов BBP будут показаны некоторые характеристики, связанные с подвижными радиомодулями MRM и с радиобазами RB.

Конструктивно модули MRM и RB почти одинаковы. Как уже говорилось, ими являются приемопередатчики, соответствующие спецификациям DECT для физического уровня. В соответствии со спецификациями DECT, радиомодули работают в полосе частот 1880 - 1900 МГц на десяти разнесенных каналах с интервалами в 1.728 МГц.

Обычно модули могут мгновенно передавать мощность около 250 мВт с циклом предусматриваемой активности согласно стандарту DECT между 4 и 96%.

Модули могут передавать сигналы, модулированные в соответствии с фильтрованной гауссовой частотной манипуляцией, которая является некогерентным вариантом гауссовой манипуляции с минимальным сдвигом, в которой BT=0.5 (BT - это произведение ширины полосы B используемого фильтра и длительности T отдельного символа).

Радиосвязь между модулями MRM и радиобазами RM происходит в соответствии с гибридной системой временного и частотного уплотнения каналов (TDM/FDM) с двойными симплексными и дуплексными соединениями.

Передача происходит во временных циклах или кадрах, имеющих длительность d (например) 10 мс, разделенных (например) на 24 временных сегмента, у которых, в соответствии со спецификациями DECT, первая половина (12) обычно служит для передач от радиобаз RB портативным радиомодулям MRM, а вторая половина (12) для передач в противоположном направлении.

Фиг. 3 показывает имеющуюся решетку временных интервалов (240) с десятью каналами для каждого кадра. В решетке время t указывается на абсциссе, а частота f_1 - на ординате. Частоты, связанные с десятью каналами, указываются от f до f_{10}, а временные сегменты, на которые разделяется каждый отдельный кадр, нумеруются от 1 до 24.

Для кадров, каждый из которых имеет длительность 10 мс, разделенную на 24 временных сегмента, каждый временной сегмент имеет длительность 416,667 мкс, 364,667 мкс, из которой могут использоваться для пакета данных, а 51 мкс - в качестве временного промежутка (защитный интервал).

Удобно, что дуплексная связь с временным делением (TDD) используется для дуплексных соединений, а сегменты на всех частотах используются для множественных соединений.

Следовательно, радиомодули MRN и RB требуют перенастройки между двумя каналами на противоположных концах частотной полосы и переключения между передачей и приемом во временном промежутке (защитном интервале) между двумя временными сегментами.

Принимающая часть радиомодулей MRM и радиобаз RB имеет супергетеродинную архитектуру с одним каскадом преобразования.

Как следует на фиг. 1, каждая радиобаза RB имеет соответствующую антенну A, а каждый из подвижных радиомодулей MRM абонентских станций имеет две антенны A1 и A2 для получения пространственного разнесения, позволяющего улучшить качество радиосоединений.

В варианте осуществления, показанном на фиг. 2, каждое интерфейсное устройство LM, связанное с каждым терминалом ввода данных, содержит главный микропроцессор 50 и процессор сигналов 51.

Главный микропроцессор 50, который образован, например, устройством V40, производимым компанией "Ниппен Илектрик" (Nippon Electric Company), может общаться с

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шиной DB связанного с ним терминала ввода данных посредством двухпортовой памяти с произвольной выборкой 52 и с другим микропроцессором 51 посредством другой двухпортовой памяти с произвольной выборкой 53.

Микропроцессор 50 связывается с памятью программ 54, например, типа стираемой программируемой постоянной памяти, и с буферной памятью с произвольной выборкой 55 для данных.

Микропроцессор 50 и память 55 связываются с устройством 56 для управления сопряжением с памятью и декодирования портов ввода-вывода. Это устройство формируется как интегральная схема ASIC (интегральная схема для специальных приложений) с высоким уровнем интеграции.

Микропроцессор 51 является устройством для обработки цифровых сигналов, например, устройством TMS320, изготовляемым компанией "Тексас Инструмент", и программируется для управления низкоуровневыми функциями управления доступом к среде (МАС), такими, как форматирование и деформатирование кадров и сегментов, синхронизация сегментов и кадров, обнаружение ошибок, сканирование каналов связи и так далее.

Процессор 51 соединяется также с устройством 57, которое извлекает тактовые сигналы из сигналов, принимаемых подвижным радиомодулем MRM, и генерирует синхронизирующие сигналы, а также осуществляет любое кодирование для защиты передаваемых данных. Устройство 57 может также изготавливаться в виде одиночной интегральной схемы ASIC для специальных приложений.

Указанное устройство связывается с буфером 58, который действует в качестве защитной защелки. Процессор 51 связывается посредством буфера и многожильного кабеля СС с устройством 59 в подвижном радиомодуле MRM для управления радиосхемами передачи и приема 60. Устройство 59 также может производиться в виде специализированной интегральной схемы ASIC.

Удобно, что устройство LM запитывается от шины DB терминала ввода данных, например, посредством двух проводников, обозначенных цифровой позицией 60 на фиг. 2. Кроме того, подвижный радиомодуль MRM запитывают от источника электрического питания адаптерного устройства LM, например, посредством двух проводников, обозначенных цифровой позицией 61 на фиг. 2, которые проходят через многожильный соединительный кабель СС.

Как упоминалось выше, с аппаратурной точки зрения, процессоры групповых сигналов BBP устройства концентратора МС имеют ту же структуру, что и логические модули LM, вводимые в терминалы ввода данных абонентских станций Т. Фактически большинство функций процессоров групповых сигналов соответствует функциям, выполняемым модулями LM. Эти функции включают в себя, в частности:

- создание и ликвидацию сегментных структур,
- создание и ликвидацию логических каналов,
- контроль за свободными каналами во входящих коммуникациях,
- распространение сообщений "без соединения" и системы персонального вызова,
- передача обслуживания между логическим и "межэлементным" каналами,
- управление быстрыми процедурами для обнаружения и исправления ошибок.

Интерфейсные адаптеры LM терминалов ввода данных предусматривают также выполнение следующих функций:

- создание и обновление карты использования физических каналов связи и выбор канала для каждого соединения, которое должно устанавливаться, и
- решение осуществить либо внутриэлементную, либо межэлементную передачу обслуживания и ее инициирование.

Адаптерные модули LM действуют также в качестве интерфейсов между режимом DECT и прикладным окружением соответствующих терминалов ввода данных. Модуль LM таким образом соответствует сетевой операционной системе (администратору локальной сети), находящейся в терминале ввода данных, точно таким же образом, как адаптер сети Ethernet, посредством стандартного интерфейса "Спецификаций интерфейсов сетевых драйверов фирмы "Майкрософт" (Microsoft Network Driver Interface Specification).

Два решающих требования для применения спецификаций DECT в локальной сети LAN - это необходимость использования с максимальной эффективностью спектральных ресурсов и необходимость минимизировать задержку, вносимую DECT. Для достижения обеих этих целей необходимо использовать специальные протоколы.

Так как поток данных характеризуется короткими транзакциями, располагаемыми между продолжительными паузами, представляется невозможным сохранить соединения между станциями пользователей и радиобазами, постоянно открытыми ввиду их существенного недоиспользования. Поэтому радиосоединения в сети устанавливаются только тогда, когда есть данные для передачи, и прерываются при отсутствии последних с тем, чтобы освободить радиоканалы для использования другими пользователями.

С этой целью главный процессор 50 каждого модуля LM программируется для работы следующим образом.

Каждый раз, когда данные подводятся к буферной памяти 55 для передачи посредством связанного подвижного радиомодуля MRM, главный микропроцессор 50 устанавливает радиосоединение посредством микропроцессора 51 (с радиобазой, определенной ниже, и с использованием сегментов канала или частоты, определенных ниже). Радиосоединение, установленное таким образом, сохраняется на протяжении всего времени, необходимого для передачи данных в память 55. После передачи данных радиосоединение не закрывается сразу же, а сохраняется в течение предварительно определенного периода времени. Главный микропроцессор 50 используется для обработки краткосрочной статистики,

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относящейся к трафику связи терминала ввода данных (например, на период в полчаса или час). Затем радиосоединение, обеспечивающее передачу данных, прерывается с задержкой после момента окончания передачи данных, причем задержка определяется адаптивно на основе среднего трафика, который действовал на терминал ввода данных. Это уменьшает ненужные паузы, так как, в большинстве случаев, не обязательно заново устанавливать радиосоединение, когда для передачи поступает последующий поток данных.

Для того чтобы выбрать радиобазу, с которой устанавливается соединение, каждый адаптерный модуль LM станции пользователя работает следующим образом.

В соответствии со стандартом DECT главный микропроцессор 50 адаптера (LM) каждой абонентской станции используется циклически для сканирования всех сегментов всех каналов посредством связанного подвижного радиомодуля MRM для того, чтобы определить уровень сигнала, излучаемого каждой неподвижной радиобазой RB в каждом сегменте для каждого канала или частоты. На основе оценки уровней сигналов, определяемых таким образом, микропроцессор 50 может установить, какая из неподвижных радиобаз RB является ближайшей. Во время сканирования процессор используется также для декодирования сигналов, указывающих для каждого сегмента радиобазу RB, которая может быть активна.

Благодаря такому "картографированию" для передачи данных главный процессор 50 устройства LM каждого терминала пользователя может выбирать ближайшую радиобазу, у которой не все сегменты заняты в рассматриваемый момент времени.

Подобная процедура позволяет избежать тщетных попыток установить радиосоединение с радиобазой, которая, хотя в является ближайшей, полностью занята в рассматриваемый момент.

В соответствии со стандартом DECT процессоры групповых сигналов BBP устройства концентратора MC используются для циклического сканирования каналов или частот f_1-f_{10} посредством связанных радиобаз RB. В частности, сканирование происходит синхронно с циклическим сканированием, осуществляемым устройствами LM терминалов пользователей. Кроме того, главные процессоры 50 модулей интерфейсных адаптеров LM используются для проведения сканирования на один канал вперед. Другими словами, если в ходе их сканирования неподвижные радиобазы RB "опрашивают" канал или частоту f_1, в тот же момент подвижные радиомодули "опрашивают" канал или частоту f_{i+1}.

Все это позволяет минимизировать время, необходимое для установления радиосоединения между терминалом пользователя и неподвижной радиобазой.

Главные процессоры 50 интерфейсных адаптеров LM абонентских станций и процессоры групповых сигналов BBP концентратора MC используются для выполнения процедур асимметричного соединения и соединения с множеством однонаправленных каналов передачи данных стандарта DECT для определения, в каком сегменте передавать.

Процедура с множеством однонаправленных каналов обеспечивает одновременное присвоение соединению, связанному с одной абонентской станцией, нескольких сегментов (однонаправленных каналов). Ширина полосы для абонентской станции, таким образом, может увеличиваться, например, от дуплекса с 32 кб/с (один однонаправленный канал) до (теоретически), например, дуплекса с 384 кб/с с использованием всех двенадцати пар сегментов (12 однонаправленных каналов).

Так как трафик в локальной сети обычно очень асимметричен, в частности, при необходимости иметь значительные ширины полос, доступные в одном направлении, спецификации DECT включают механизмы, которые обеспечивают использование в одном направлении верхних и нижних сегментов соединения. Соединение этого типа должно образовывать часть соединения с множеством однонаправленных каналов, в которой по меньшей мере одно другое соединение остается дуплексным для обеспечения маршрута для управляющих данных в противоположном направлении. Результат состоит в том, что пользователь может обращаться почти ко всей ширине полосы (352 кб/с) путем занятия половины сегментов, как показано на фиг. 4, что связано с асимметричным соединением с множеством однонаправленных каналов (5, 1).

Наконец, программное обеспечение, используемое в локальной сети LAN, включает в себя процедуры обнаружения и исправления ошибок в соответствии со спецификациями DECT. Спецификации предусматривают на уровне 2 (MAC/DLC) некоторые механизмы, которые были разработаны соответственно для этой цели и главные характеристики которых таковы:

- управление доступом к среде (MAC) обеспечивает услуги, определяемые как "I_p" (защищенный информационный канал), с пропускной способностью 25.6 кб/с на соединение и коэффициентом ошибки 10^{-5}; эти услуги основаны на механизме повторной передачи, который отличается высоким быстродействием и простотой благодаря использованию однооконного пакета;
- DLC (управление каналом передачи данных) обеспечивает услуги, определенные как "кадровое реле", которое защищают данные от любых ошибок, вносимых во время изменений передачи обслуживания и соединения и от остаточных ошибок канала I_p.

В заключение необходимо отметить, что, хотя принцип изобретения остается неизменным, варианты осуществления и детали конструкции могут широко варьироваться по сравнению с описанными и проиллюстрированными только в рамках не вносящего ограничений примера, без изменения объема настоящего изобретения.

Формула изобретения:

1. Способ обмена данными между множеством абонентских станций по бескабельной локальной сети через центральную управляющую станцию, производимого с применением стандарта связи DECT, использующего широкую полосу

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частот, разделенную на множество (n) каналов (f_1-f_{10}) и на заданное число временных сегментов (2m), отличающийся тем, что периодически транслируют с центральной управляющей станции (C) сигналы для каждого из множества (n) каналов (f_1-f_{10}) и для каждого из временных сегментов (2m), периодически сканируют множеством абонентских станций (T) указанные сигналы для определения уровня сигнала каждого из множества (n) каналов (f_1-f_{10}) и соответственно определения занятых или незанятых временных сегментов (2m), формируют и периодически корректируют с помощью каждой из абонентских станций (T) список значений уровней сигналов и незанятых временных сегментов (2m), устанавливают радиосвязь между одной из выбранных абонентских станций (T) и центральной управляющей станцией (C) в случае, когда выбранная абонентская станция (T) содержит информацию, подлежащую передаче, производят обмен информацией между указанной выбранной абонентской станцией (T) и центральной управляющей станцией (C).

2. Способ по п. 1, отличающийся тем, что поддерживают радиосвязь в течение заданного периода времени до завершения обмена информацией.

3. Способ по п. 1, отличающийся тем, что поддерживают радиосвязь в течение адаптивно заданного периода времени на основе статистических данных связной нагрузки относительно выбранной абонентской станции (T), полученных в заранее определенный период времени.

4. Способ по п. 1, отличающийся тем, что обеспечивают асимметричную многоканальную широкополосную связь с использованием множества сегментов из заданного числа временных сегментов (2m), используемых одновременно при установлении указанной радиосвязи.

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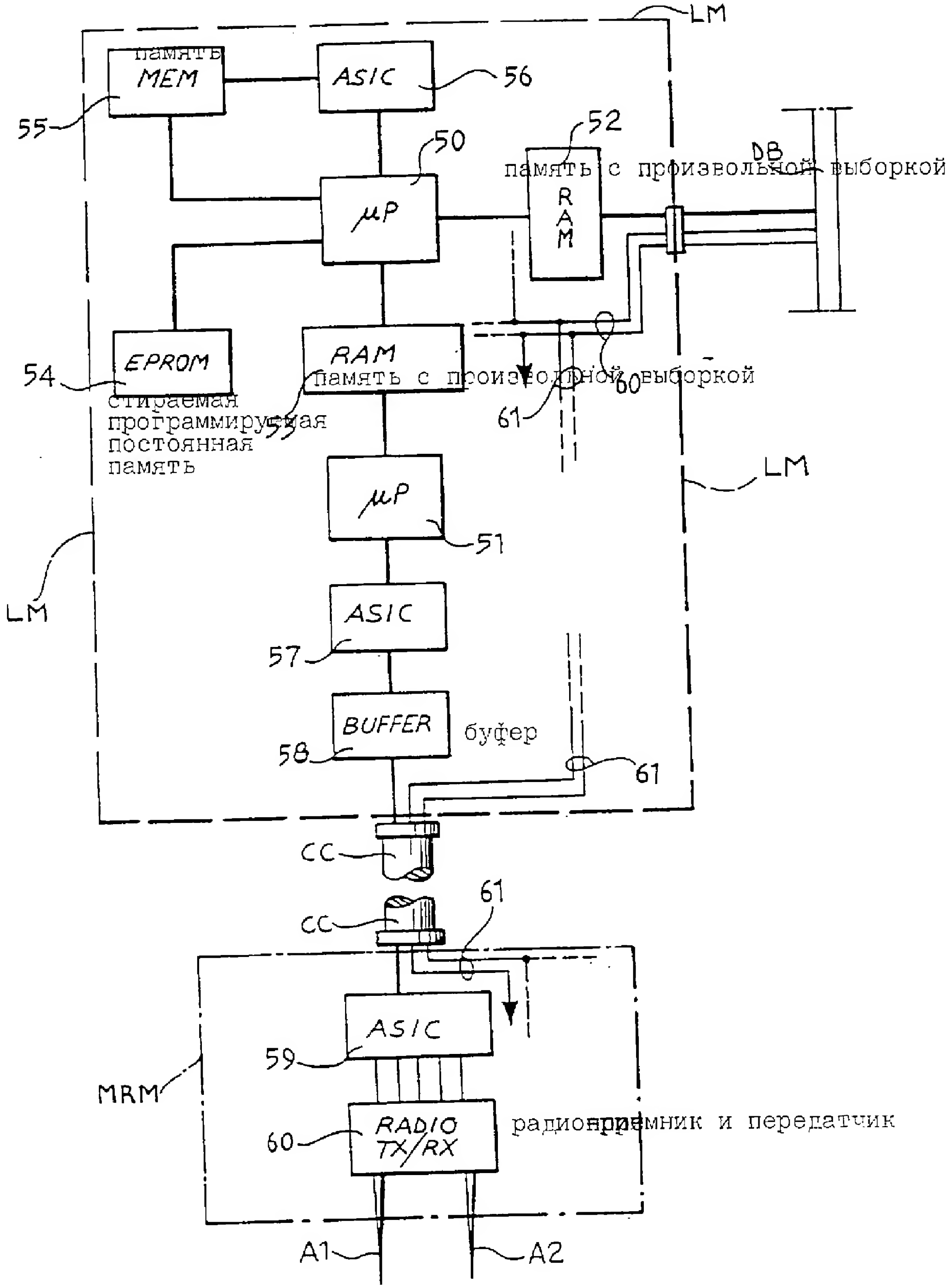

Фиг. 2

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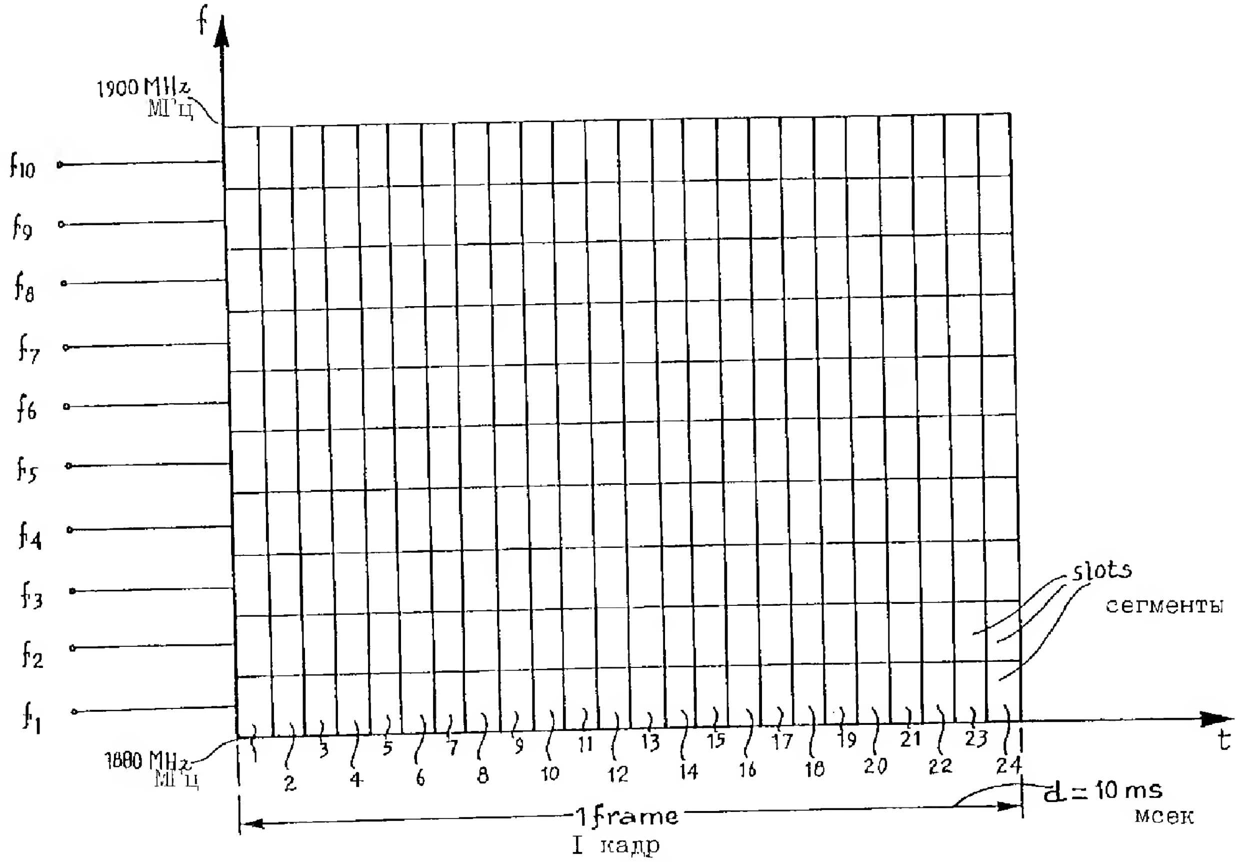
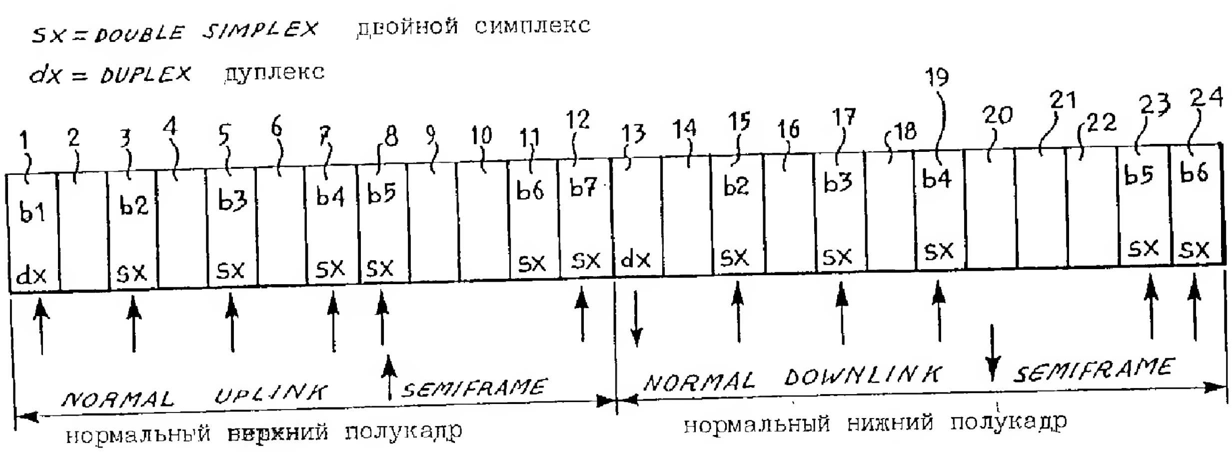


Фиг. 3



Фиг. 4

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(71) Applicant *(for all designated States except US):* SIXTEL S.P.A. [IT/IT]; Via Jervis, 77, I-10015 Ivrea (IT).

(72) Inventors; and
(75) Inventors/Applicants *(for US only)* : BUD, Andrew [GB/IT]; Via Miniere, 11, I-10015 Ivrea (IT). FURNO, Rino [IT/IT]; Via Portici, 24, I-13040 Zimone (IT).

(74) Agents: QUINTERNO, Giuseppe et al.; Jacobacci-Casetta & Perani S.p.A., Via Alfieri, 17, I-10121 Torino (IT).

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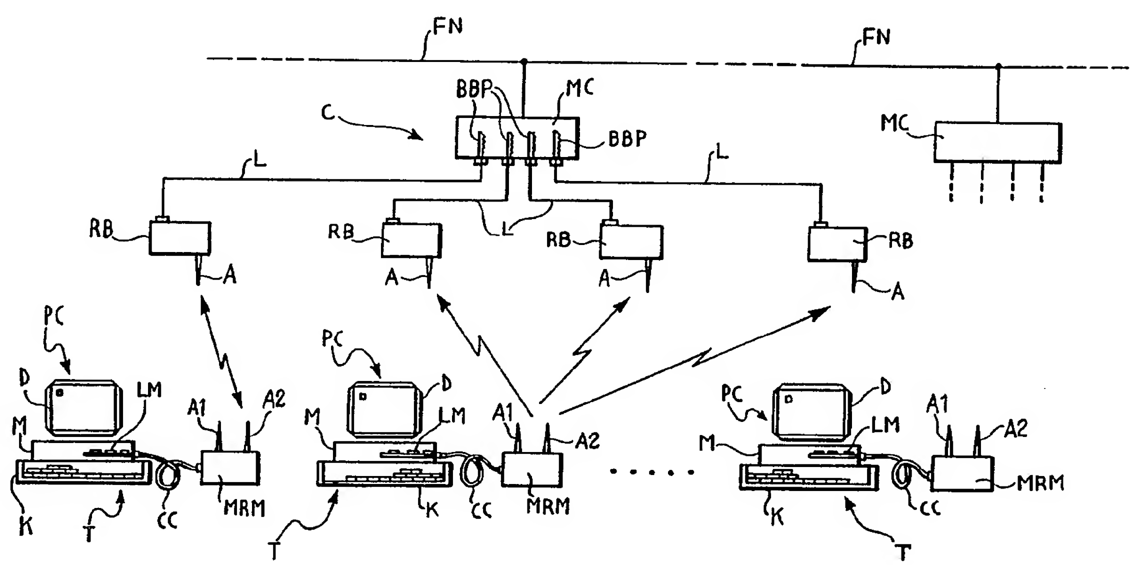

(57) Abstract

The network enables data to be communicated by radio, in accordance with the DECT standard, between the data terminals (PC) of a plurality of user stations (T), by means of a fixed central control device (C). Each user station (T) is associated with a mobile radio transmitter/receiver module (MRM) which is separate and distinct from the data-terminal (PC), and with an adaptor device (LM) which acts as an interface between the data terminal (PC) and the radio module (MRM) and which is physically incorporated in the data terminal (PC) and is connected to the radio module (MRM) by a flexible multicore cable (CC). The central control device (C) includes a multiplicity of fixed radio modules or bases (RB) and a fixed concentrator (MC) which is connected to the fixed radio bases (RB) by connecting lines (L).

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Cordless local area network (radio LAN) with central unit

The present invention relates to a local area network (a LAN) and, more specifically, to a network of the cordless (or wireless) type, particularly a network which operates in accordance with the DECT standard to enable data to be communicated by radio between a plurality of user stations each comprising a respective data terminal, by means of a fixed central control device.

Local networks have become increasingly widespread in the informatics and telematics world for short-range connections for enabling the transmission and distribution of data and services between a plurality of users within the same area, for example, in the same building. A local network enables many data terminals of different kinds, such as personal computers (PCs), minicomputers, printers, etc. to be connected in an extremely flexible manner, enabling very fast transmission speeds of the order of hundreds of thousands of Kbits/sec.

Up to now, most local networks have been of the wired type, that is, of the type in which the connections between the user stations and the central control devices are formed entirely by wires.

The appearance on the market of portable computers such as laptop personal computers, has created a need for cordless LANs.

A cordless local area network reduces installation costs because it eliminates the need to install connecting cables. This type of network can also be

formed in situations in which it would be difficult or impossible to install connecting wires, such as, in buildings which do not have sockets for LANs or in which there are architectural constraints.

A cordless LAN network may represent the ideal solution in an organisation in which the positions of the user stations or the number of stations connected in the network are subject to frequent changes or modifications.

A cordless LAN also represents the ideal solution for organisations which are subject to frequent changes of location. In this case, it would in fact be neither practical nor economical to transfer a wired LAN.

Finally, as stated above, a LAN network enables data to be communicated even by portable personal computers, without limiting the mobility of these new devices.

The network according to the invention operates in accordance with the DECT (Digital European Cordless Telecommunications) standard developed by ETSI - the European Telecommunications Standards Institute - which defines the specifications for radio connections between users and a network in a private environment.

The DECT system operates in the band between 1880 MHz and 1900 MHz and provides for radio transmission by means of a hybrid time and frequency multiplex system.

The characteristics of the DECT standard are described, for example, in "Digital European Cordless Telecommunications Services and Facilities", ETSI DR/RES 3003, June 1991 and in "Data Services in DECT",

A. Bud, Fifth International IEE Conference on Land Mobile Radio, Warwick, December 1989.

The cordless local area network according to the invention is characterised in that the data terminal of each user station is associated with:

- a mobile radio transmitter/receiver module which is separate and distinct from the data-terminal, and

- a microprocessor adaptor device for acting as an interface between the data terminal and the associated mobile radio module, the adaptor device being incorporated physically in the data terminal and connected to the mobile radio module by a flexible multicore cable,

and in that the central control device includes:

- a multiplicity of radio modules or bases for installation in respective predetermined fixed positions and for transmitting/receiving packets of data to/from the radio module of one or more user stations, and

- a microprocessor concentrator (hub) which is intended to be installed in a fixed position and to be connected to the fixed radio bases and which is programmed to control the communications between the user stations by means of the radio bases, according to predetermined procedures and protocols.

Typically, the data-terminals of the user stations may, for example, be personal computers and the microprocessor adaptor device is conveniently produced

in the form of a "half-size"-format card or daughter board incorporated in the PC and connected to the bus thereof. The electrical supply for the adaptor is thus conveniently derived from the data-terminal bus.

Moreover, the transmitter/receiver radio module to advantage takes its electrical supply from the associated adaptor board by means of conductors which extend through the flexible multicore cable connecting it to the board.

To advantage, each user station radio transmitter/receiver module has two omnidirectional antennae for achieving space "diversity" to improve the characteristics of the radio connection with the fixed radio modules or bases.

Conveniently, but not necessarily, the fixed central control device may be arranged for connection to a fixed network, for example, an Ethernet network or a Token Ring or RS232 network.

Further characteristics and advantages of the invention will become clear from the following detailed description of a cordless LAN network operating in accordance with the DECT standard, the description being given with reference to the appended drawings, provided purely by way of non-limiting example, in which:

Figure 1 is a block diagram of the LAN network,

Figure 2 is a circuit diagram showing the structure of an adaptor and a mobile radio module associated with each data-terminal of the LAN network shown in Figure

1,

Figure 3 is a time/frequency diagram relating to the manner in which radio transmission is effected according to a hybrid TDM/FDM system in the LAN network of Figure 1, and

Figure 4 shows an example of a frame for an asymmetric multi-bearer connection which can be formed in the LAN network of Figure 1.

With reference to Figure 1, a cordless local area network LAN formed in accordance with the specifications of the DECT standard includes a plurality of user stations T and a fixed central control device, generally indicated C.

Each user station T includes a respective data terminal which, in general, may be constituted by any device, such as a processor, a printer, etc., which can send and/or receive digital data by means of a communications network. In the embodiment shown by way of example in Figure 1, the data terminals of the user stations T are constituted by personal computers PC having standard network and applications software of the LAN Manager type. The personal computers may, for example, be Olivetti 1/D33 devices, each including a keyboard K, a display screen D and a processing module M.

Each data terminal PC is connected to a respective mobile radio transmitter/receiver (transceiver) module indicated MRM, of a type conforming to the DECT specifications for the physical layer.

The processing module M of each data-terminal PC incorporates a respective microprocessor adaptor device, indicated LM. The microprocessor adaptor is suitable for acting as an interface between the respective data terminal and the associated mobile radio module MRM. For this purpose, as shown schematically in Figure 2, the microprocessor adaptor LM is connected to the data bus DB of the processing module M of the data terminal. The adaptor LM is also connected to the mobile radio module MRM associated with the data terminal by means of a multicore flexible cable CC (Figure 1 and 2).

The central control device C includes a multiplicity of fixed radio modules or bases FRM installed in respective predetermined fixed positions for transmitting/receiving packets of data to/from the mobile radio module MRM of one or more user stations T.

The radio bases RB are connected, for example, by electrical wires L, to a microprocessor concentrator MC which is installed in a fixed position and is programmed to control the communications between the user stations T by predetermined procedures and protocols, in accordance with the DECT standard, by means of radio connections established between the mobile radio modules MRM and the radio bases RB.

Preferably, but not necessarily, the concentrator MC may be arranged for connection to a fixed network FN, for example an Ethernet network or a Token Ring or RS 232 network. Concentrators MC of other local networks LAN may possibly be connected to the fixed network.

The integrated system described with reference to

Figure 1 can perform the function of an MAC (medium access control) level multi-port bridge to enable the user stations T to transmit and receive packets of data which are packaged in accordance with the DECT standard format and are exchanged by radio, by means of the fixed portion C of the system. This portion acts as a very rapid packet-switching system and directs the packets received towards the destination user stations or towards the wired network FN.

The system described operates in accordance with the DECT standard. The DECT standard connection between the user stations T and the fixed portion C of the system replaces only the MAC level of the Ethernet system.

By virtue of the lines L, the radio bases RB can be installed up to distances of the order of 100 metres from the concentrator MC. By carrying out functions, such as connection handover, which are provided for in the DECT standard, almost complete continuity of service between the two or more radio bases RB used can be established.

The concentrator MC may be constituted, for example, by an Olivetti M300 personal computer with an Intel 386Sx processor operating at 16 MHz.

This concentrator incorporates baseband processors BBP connected in an orderly manner to respective associated radio bases RB.

Conveniently, the baseband processors BBP of the concentrator MC and the interface adaptors LM of the user station T may be in the form of half-size format

PC circuit boards and, in practice, may conveniently have the same structure at the hardware level and be differentiated only at the software level. The structure of an interface adaptor LM of a user station will be described in greater detail below with reference to Figure 2.

The concentrator MC as a whole is responsible for controlling the entire system and, in particular:

- the functioning of the high levels of the DECT protocols,

- the control of the various resources of the network,

- the switching of the packets of data, and possibly

- the interfacing between the cordless network LAN and the wired network FN.

The high levels of the DECT protocols provide for services such as fast handover, user authentication and the creation of virtual connections which enable physical connections to be established without massive exchanges of data.

Before the merits of the structure of the functions of the LM devices and of the band base processors BBP are discussed further, some characteristics relating to the mobile radio modules MRM and to the radio bases RB will be set out.

Structurally, the modules MRM and RB are almost identical. As already stated they are transceivers conforming to the DECT specifications for the Physical

Layer. In accordance with the DECT specifications, the radio modules operate in the band between 1880 MHz and 1900 MHz on ten channels spaced at 1.728 MHz intervals.

Typically, the modules can instantaneously transmit a power of about 250 mW with an envisaged activity cycle according to the DECT standard of between 4% and 96%.

The modules can transmit signals modulated according to filtered Gaussian FSK which is a non-coherent version of GMSK in which BT = 0.5 (BT is the product of the bandwidth B of the filter used and the duration T of the individual symbol).

Radio communications between the MRM modules and the radio bases RM take place according to a hybrid time and frequency multiplex system (TDM/FDM) with double simplex and duplex connections.

Transmission takes place within time cycles or frames having durations d of (for example) 10 ms, divided (for example) into 24 slots, of which, in accordance with the DECT specifications, a first half (12) normally serve for transmissions from the radio bases RB to the portable radio modules MRM and the second half (12) for transmissions in the opposite direction.

Figure 3 shows the grid of the slots (240) available with ten channels for each frame. In the grid, the time t is indicated on the abscissa and the frequency f is indicated on the ordinate. The frequencies associated with the ten channels are indicated f_1-f_{10} and the slots into which each individual frame is divided are numbered 1-24.

With frames each of 10 ms divided into 24 slots, each slot has a duration of 416.667 μs of which 364.667 μs can be used for a packet of data and 51 μs as a time interval (a guard space).

Conveniently, a time-division duplex (TDD) is used for duplex connections and slots at all the frequencies are used for multiple connections.

The radio modules MRM and RB therefore need to be able to retune themselves between two channels at opposite ends of the band and to switch between transmission and reception within the time interval (the guard space) between two slots.

The receiving portions of the radio modules MRM and of the radio bases RB conveniently have superheterodyne architecture with a single conversion stage.

As is clear from Figure 1, each radio base RB has a respective antenna A and the mobile radio modules MRM of the user stations each have two antennae A1 and A2 for achieving space diversity in order to improve the quality of the radio connections.

In the embodiment shown in Figure 2, each interface device LM associated with each data terminal includes a main microprocessor 50 and a signal processor 51.

The main microprocessor 50 which is constituted, for example, by a V40 device produced by Nippon Electric Company, can converse with the bus DB of the associated data terminal by means of a dual-port RAM memory 52 and with the other microprocessor 51 by means of another dual-port RAM memory 53.

The microprocessor 50 is associated with a program memory 54, for example, of the EPROM type and a RAM buffer memory 55 for the data.

The microprocessor 50 and the memory 55 are associated with a device 56 for controlling the interfacing with the memory and decoding the I/O ports. This device is conveniently formed as a large-scale integration ASIC integrated circuit (an application-specific integrated circuit).

The microprocessor 51 is a device for processing digital signals, for example, a TMS320 device produced by Texas Instruments and is programmed to control low-level MAC functions such as the formatting and deformatting of the frames and of the slots, the synchronisation of slots and frames, the detection of errors, the scanning of the communication channels, etc.

The processor 51 is also connected to a device 57 which extracts the clock signals from the signals received by the mobile radio module MRM and generates the timing signals and also effects any coding for protecting the data to be transmitted. The device 57 may also conveniently be produced in the form of a single ASIC integrated circuit.

This device is associated with a buffer 58 which acts as a protection latch. The processor 51 is connected by means of the buffer and the multicore cable CC to a device 59 within the mobile radio module MRM for controlling the radio transmission/reception circuits 60. The device 59 may also conveniently be produced in the form of an ASIC integrated circuit.

Conveniently, the device LM draws its electrical supply from the bus DB of the data terminal, for example, by means of the two conductors indicated 60 in Figure 2. Moreover, the electrical supply of the mobile radio module MRM to advantage is derived from that of the adaptor device LM, for example, by means of two conductors indicated 61 in Figure 2, which extend through the multicore interconnecting cable CC.

As stated above, from a hardware point of view, the baseband processors BBP of the concentrator device MC have the same structure as the logic modules LM fitted in the data terminals of the user stations T. In fact most of the functions of the baseband processors correspond to functions carried out by the modules LM. These functions include, in particular:

- the creation and dismantling of the slot structures,

- the creation and dismantling of logic channels,

- the monitoring of the free channels in the incoming communications,

- the propagation of "connectionless" and "paging" messages,

- handover between logic and "inter-cell" channels,

- the control of rapid procedures for detecting and correcting errors.

The interface adaptors LM of the data terminals are arranged also to perform the following functions:

- the creation and updating of a map of the usage of the physical communications channels and the selection of the channel for each connection to be established, and

- the decision to effect either intra-cell or inter-cell handover and the initiation thereof.

The adaptor modules LM also act as interfaces between the DECT environments and the applications environments of the respective data terminals. The module LM thus responds to the network operating system (the LAN manager) resident in the data terminal in exactly the same manner as an Ethernet network adaptor by means of a Microsoft Network Driver Interface Specification standard interface.

Two critical requirements for the application of the DECT specifications in a local area network LAN are the need to use the spectral resources with maximum efficiency and the need to minimise the delay introduced by the DECT. In order to achieve both these objectives, it is necessary to use specific protocols.

Since the data traffic is characterised by short transactions interposed between long silences it is inconceivable to keep the connections between the user stations and the radio bases open permanently since they would be massively underused. The radio connections are therefore established in the network only when there are data to transmit and are closed during periods of inactivity in order to free radio channels for use by other users.

For this purpose, the main processor 50 of each module LM is programmed to operate in the following manner.

Each time data are admitted to the buffer memory 55 for transmission by means of the associated mobile radio module MRM, the main microprocessor 50 sets up a radio connection by means of the microprocessor 51 (with a radio base determined in the manner which will be described below and with the use of slots of a channel or frequency determined in the manner which will also be described below). The radio connection thus opened is maintained throughout the time necessary for the transmission of the data in the memory 55. After the data have been transmitted the radio connection is not closed immediately but is kept open for a predetermined period of time. Conveniently, the main microprocessor 50 is arranged to process a short-term statistic relating to the communications traffic of the data terminal (for example, over a period of half an hour or an hour). The radio connection opened for the transmission of data is then closed with a delay after the moment at which the transmission of data is completed, the delay being determined adaptively on the basis of the mean traffic which has affected the data terminal. This reduces useless periods since, in many cases, it is not necessary to reopen the radio connection when a further flow of data arrives for transmission.

In order to select the radio base with which to establish the connection, each user station adaptor module LM operates in the following manner.

In accordance with the DECT standard, the main microprocessor 50 of the adaptor (LM) of each user

station is arranged cyclically to scan all the slots of all the channels by means of the associated mobile radio module MRM in order to detect the level of the signal emitted by each fixed radio base RB in each slot for each channel or frequency. On the basis of the levels of the signals thus detected, the microprocessor 50 can establish which is the nearest fixed radio base RB. The processor is also arranged, during the scanning, to decode the signals indicative, for each slot, of the radio base RB which is (possibly) active.

By virtue of this "mapping", in order to transmit data, the main processor 50 of the device LM of each user terminal can select the nearest radio base of which not all the slots are occupied at the time in question.

This procedure avoids futile attempts to establish a radio connection with a radio base which, although it is the nearest, is fully occupied at the time in question.

In accordance with the DECT standard, the baseband processors BBP of the concentrator device MC are arranged to scan the channels or frequencies f_1 - f_{10} cyclically by means of the associated radio bases RB. In particular, the scanning takes place in synchronism with the cyclical scanning effected by the devices LM of the user terminals. Moreover, the main processors 50 of the interface adaptor modules LM are arranged to carry out the scanning one channel in advance. In other words, if, in the course of their scanning, the fixed radio bases RB are "interrogating" the channel or frequency f_i, at the same moment, the mobile radio modules are "interrogating" the channel or frequency f_{i+1}.

This minimises the time needed to establish a radio connection between a user terminal and a fixed radio base.

Conveniently, the main processors 50 of the interface adaptors LM of the user stations and the baseband processors BBP of the concentrator MC are arranged to carry out the DECT Multibearer and Asymmetric Connection procedures in order to determine in which slot to transmit.

The multibearer procedure enables several slots (bearers) to be assigned simultaneously to the connection associated with a single user station. The bandwidth available for a user station may thus be increased from, for example, 32 kb/s duplex (single bearer) up to (theoretically), for example, 384 kb/s duplex with all twelve pairs of slots (12 bearers) in use.

Since the traffic in a local area network is typically very asymmetrical with the need to have considerable bandwidths available in one direction in particular, the DECT specifications include mechanisms which enable the uplink and downlink slots of a connection to be used in a single direction. A connection of this type must form part of a multibearer connection in which at least one other connection remains duplex to provide a route for control data in the opposite direction. The result is that a user can access almost the whole of the bandwidth (352 kb/s) by occupying half of the slots as shown in Figure 4, which relates to an asymmetric multibearer connection (5, 1).

Finally, the software used in the network LAN

conveniently includes procedures for detecting and correcting errors in accordance with the DECT specifications. The specifications provide for, at the level 2 (MAC/DLC), some mechanisms which have been developed appropriately for this purpose, and the main characteristics of which are the following:

- the MAC provides a service defined as an "Ip" (a protected information channel) with a throughput of 25.6 kb/s per connection and an error factor of 10^{-5}; this service is based on a retransmission mechanism which is quick and simple by virtue of the use of a single window packet;

- the DLC (data link control) provides a service defined as "Frame Relay" which protects the data against any errors introduced during handover and connection changes and against residual errors of the Ip channel.

Naturally, the principle of the invention remaining the same, the forms of embodiment and details of construction may be varied widely with respect to those described and illustrated purely by way of non-limiting example, without thereby departing from the scope of the present invention.

CLAIMS

1. A cordless local area network (LAN) for enabling data to be communicated by radio between a plurality of user stations (T) each comprising a respective data terminal (PC), by means of a fixed central control device (C), in accordance with the DECT standard,

characterised in that the data terminal (PC) of each user station (T) is associated with:

- a mobile radio transmitter/receiver module (MRM) which is separate and distinct from the data terminal (PC), and

- a microprocessor adaptor device (LM) for acting as an interface between the data terminal (PC) and the associated mobile radio module (MRM), the adaptor being incorporated physically in the data terminal (PC) and connected to the mobile radio module (MRM) by a flexible multicore cable (CC),

and in that the central control device (C) includes:

- a multiplicity of radio modules or bases (RB) for installation in respective predetermined fixed positions and for transmitting/receiving packets of data to/from the mobile radio module (MRM) of one or more user stations (T), and

- a microprocessor concentrator (MC) which is intended to be installed in a fixed position and to be connected, by connecting lines (L), to the fixed radio bases (RB) and which is programmed to control the

communications between the user stations (T) by means of the radio bases (RB), according to predetermined procedures and protocols.

2. A local area network according to Claim 1, in which each data terminal (PC) includes a data bus (DB), the network being characterised in that the microprocessor adaptor (LM) associated with each data terminal (PC) includes:

- means (51-58) for activating/de-activating the radio connection,

- a buffer memory (55), and

- a main microprocessor (50) which is connected to the data bus (DB) of the data terminal (PC), to the buffer memory (55), and to the means (51-58) for activating/de-activating the radio connection, the main microprocessor (50) being arranged:

to control the exchange of data with the data terminal (PC) in a predetermined manner,

to admit to the buffer memory (55) the data which are to be transmitted by the associated mobile radio module (MRM), and

to pilot the activating/de-activating means (51-58) in a manner such as to activate a radio connection each time data are stored in the memory (55) and to keep the radio connection open for a predetermined period of time after the transmission of the data in the memory (55) has been completed.

3. A local area network according to Claim 2, characterised in that the main microprocessor (50) is arranged to pilot the activating/de-activating means (51, 57, 58) in a manner such that, upon completion of the transmission of the data in the memory (55), the radio link is kept open for a period of time which is determined adaptively on the basis of a communications traffic statistic relating to the data terminal (PC) and calculated over a predetermined period of time.

4. A local area network according to any one of the preceding claims in which, in accordance with the DECT standard, the radio communications between the mobile radio modules (MRM) and the fixed radio bases (RB) take place according to a mixed time and frequency multiplex system (TDM, FDM) on n channels or frequencies (f_1 - f_{10}) within a predetermined band with time cycles (frames) of predetermined duration, divided into a predetermined number (2m) of time slots, and in which the main microprocessor (50) of the adaptor (LM) of each user station (T) is arranged:

- to scan all the 2m x n slots of all the n channels (f_1 - f_{10}) cyclically by means of the associated mobile radio module (MRM) and to detect the level of the signal emitted by each fixed radio base (RB) in each slot for each channel or frequency and thus to determine which radio base (RB) is nearest the user station (T),

the network being characterised in that the main microprocessor (50) is also arranged, during the scanning, to decode the signals indicative of the radio base (RB) which is (possibly) active in each slot and to select - in order to transmit data - the nearest

radio base (RB) for which not all the slots are occupied.

5. A local area network according to any one of the preceding claims, characterised in that the concentrator device (MC) includes a multiplicity of baseband processors (BBP) each of which is associated with and connected to a respective fixed radio base (RB) and is arranged to perform the functions up to level 2 of the hierarchy of DECT protocols.

6. A local area network according to Claim 5, characterised in that the baseband processors (BBP) are arranged to scan the transmission channels or frequencies ($f_1 - f_{10}$) cyclically, in accordance with a predetermined sequence, by means of the associated radio bases (RB), and in that the main processors (50) of the adaptors (LM) of the user stations (T) are arranged to scan the transmission channels or frequencies ($f_1 - f_{10}$) in synchronism with the baseband processors (BBP) but one channel in advance thereof.

7. A local area network according to any one of Claims 2 to 6, characterised in that the main processors (50) of the adaptors (LM) of the user stations (T) and the baseband processors (BBP) of the concentrator (MC) are arranged to effect the DECT multibearer and asymmetric connection procedures in order to determine the slots in which to transmit.

8. A local area network according to any one of the preceding claims, characterised in that the adaptor (LM) of each user station (T) is formed on a half-size format PC-AT circuit board.

9. A local area network according to one of Claims 5 to 8, characterised in that the baseband processors (BBP) are incorporated in the concentrator device (MC) and are supplied thereby.

10. A local area network according to any one of the preceding claims, characterised in that each mobile radio module (MRM) receives its electrical supply from the associated adaptor (LM) by means of the multicore cable (CC) which interconnects them.

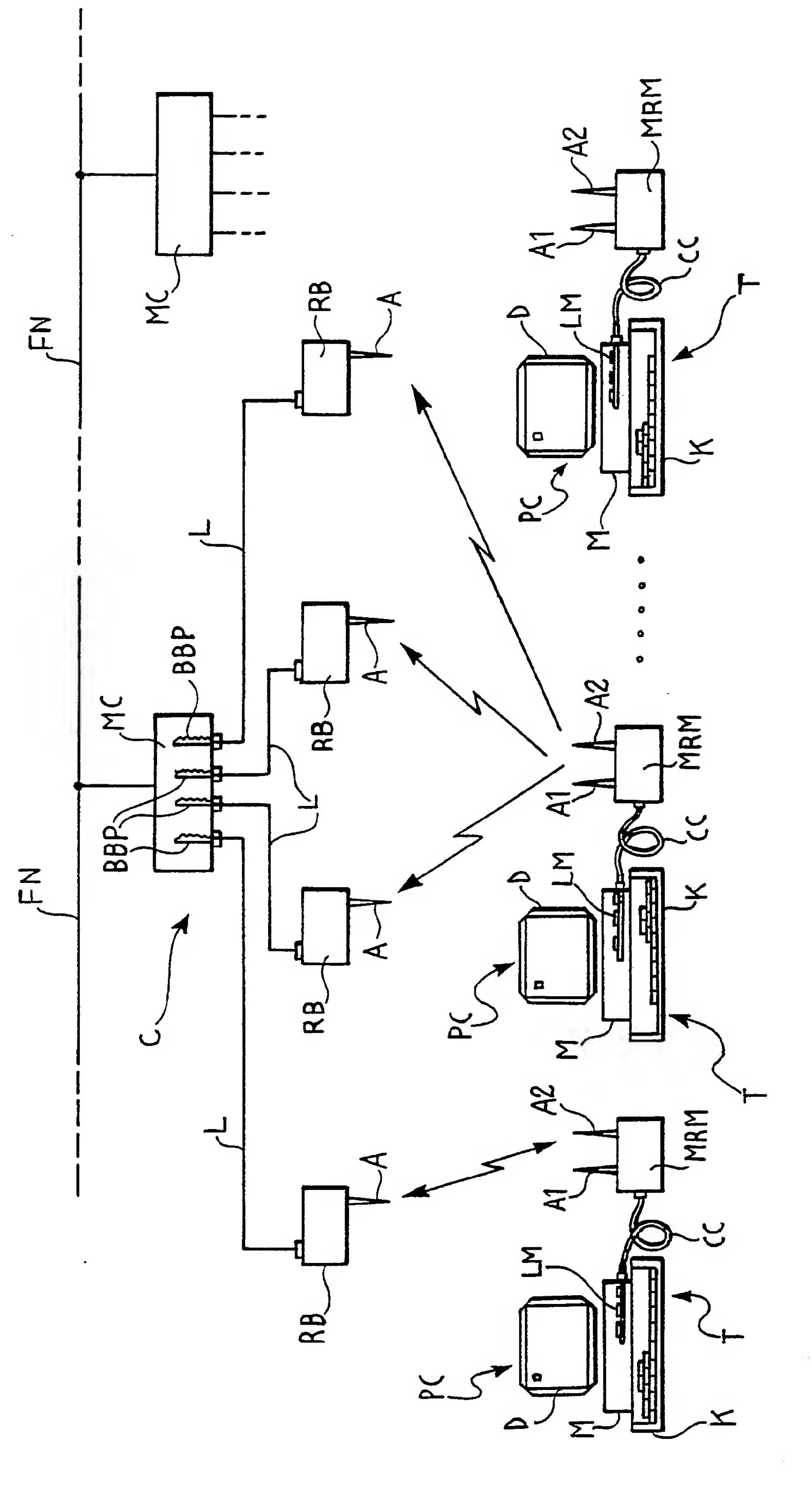
11. A local area network according to any one of the preceding claims, characterised in that each mobile radio module (MRM) has a pair of antennae (A1, A2) for achieving space "diversity".

12. A local area network according to Claim 11, characterised in that each fixed radio base (RB) has a single antenna (A).

13. A local area network according to any one of the preceding claims, characterised in that the concentrator (MC) can be connected to a fixed network (FN) such as an Ethernet or Token Ring network and can converse therewith.

14. A local area network according to Claims 1 and 5, characterised in that the adaptors (LM) of the user stations (T) and the baseband processors (BBP) of the concentrator (MC) are formed by circuit boards which are identical from the hardware point of view but which are differentiated at the software level.

1/4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MC
FN
MC
BBP
BBP
BBP
C
L
L
L
RB
A
RB
A
RB
A
RB
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A2
MRM
CC
LM
D
PC
M
K
T
A1
A2
MRM
CC
LM
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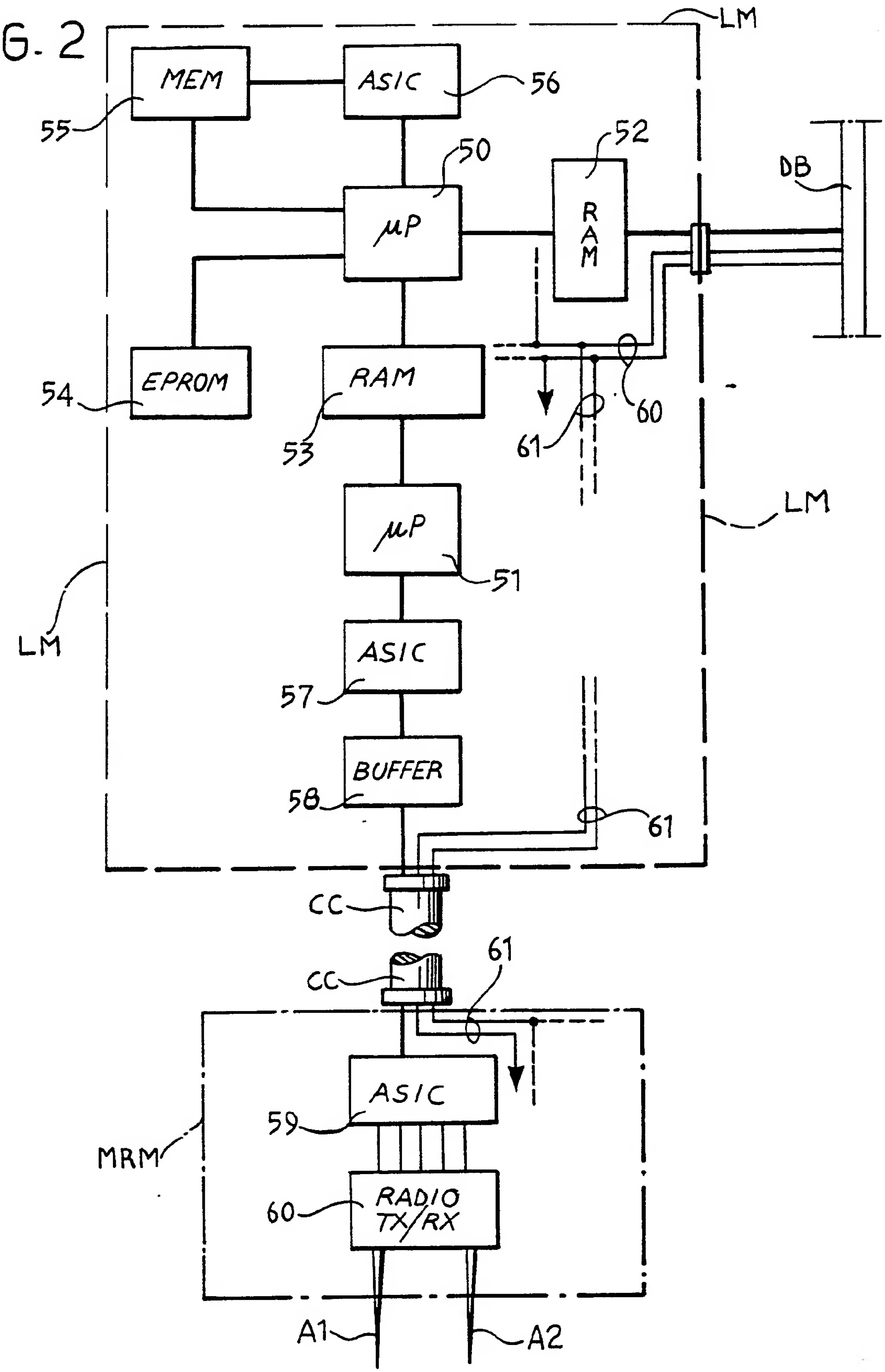
K
T

FIG. 1

FIG. 2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MEM
ASIC
56
55
52
50
DB
R
A
M
μP
EPROM
RAM
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60
53
61
LM
μP
51
LM
ASIC
57
BUFFER
58
61
CC
61
CC
ASIC
59
MRM
RADIO
TX/RX
60
A1
A2

FIG. 3

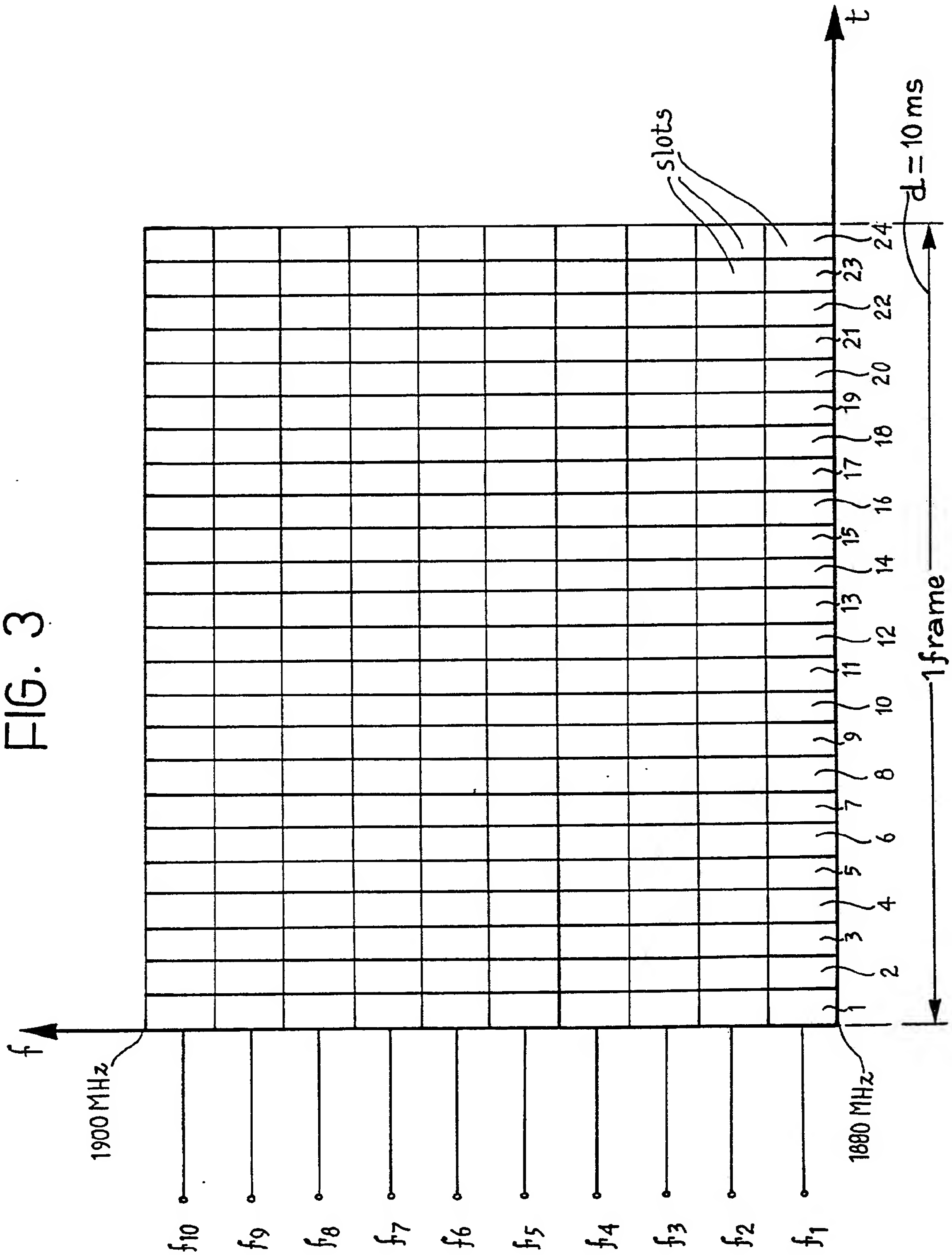
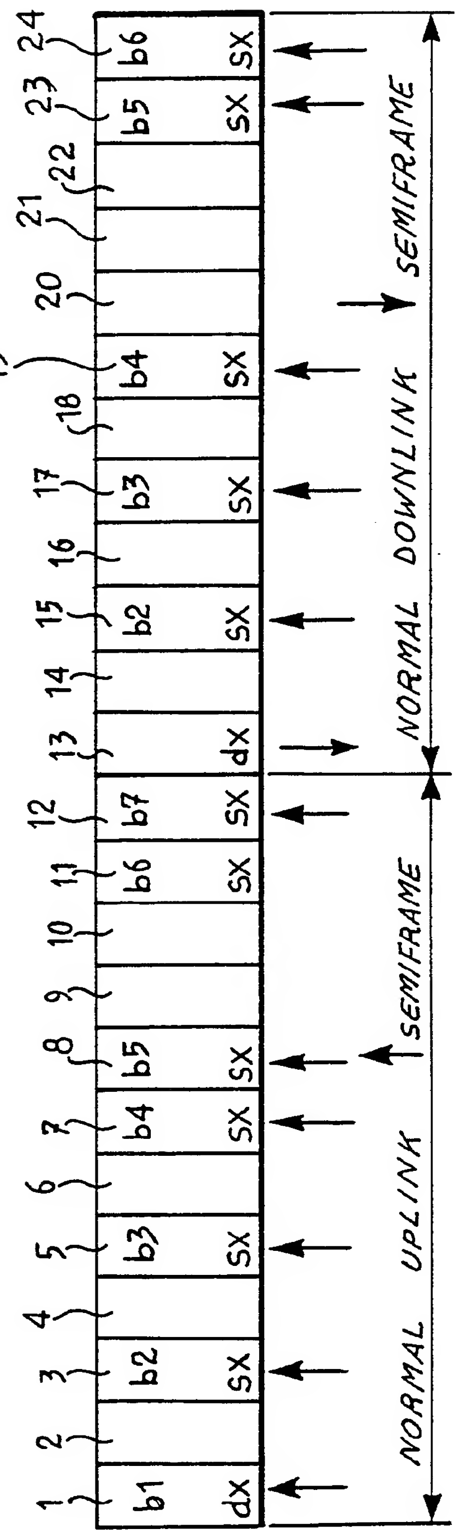

t
f
slots
d = 10 ms
1 frame
1900 MHz
1880 MHz
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18
19
20
21
22
23
24
f_{10}
f_9
f_8
f_7
f_6
f_5
f_4
f_3
f_2
f_1

FIG. 4

Sx = DOUBLE SIMPLEX

dx = DUPLEX



INTERNATIONAL SEARCH REPORT

International Application No PCT/EP 92/02230

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all) [6]

According to International Patent Classification (IPC) or to both National Classification and IPC

IPC^5: H 04 B 7/24,H 04 H 3/00,H 04 L 12/44

II. FIELDS SEARCHED

Minimum Documentation Searched [7]

Classification System	Classification Symbols
IPC^5	H 04 B 1/00,H 04 B 9/00,H 04 J 3/00,H 04 L 11/00, H 04 L 12/00,H 04 N 5/00,H 04 Q 7/00

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched [8]

III. DOCUMENTS CONSIDERED TO BE RELEVANT [9]

Category *	Citation of Document, [11] with indication, where appropriate, of the relevant passages [12]	Relevant to Claim No. [13]
E	US, A, 5 079 628 (TOMIKAWA) 07 January 1992 (07.01.92),	1,11, 12
P,A	see abstract; fig. 9; claims 1,10,15,18-20.	2-10, 13-14
	--	
X	EP, A2, 0 257 947 (ATT) 02 March 1988 (02.03.88),	1,11, 12
A	see abstract; fig. 1; claims 1-4,7,8.	2-10, 13-14
	--	
A	US, A, 4 665 519 (KIRCHNER et al.) 12 May 1987 (12.05.87), see abstract; fig. 1-3; claims 1-8.	1-14
	--	
A	GB, A, 2 125 257 (PLESSEY) 29 February 1984 (29.02.84), see abstract; claims 1-3.	1-14

* Special categories of cited documents: [10]

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

IV. CERTIFICATION

Date of the Actual Completion of the International Search	Date of Mailing of this International Search Report
28 December 1992	15 JAN 1993
International Searching Authority	**Signature of Authorized Officer**
EUROPEAN PATENT OFFICE	BLASL e.h.

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)

Category *	Citation of Document, [11] with indication, where appropriate, of the relevant passages	Relevant to Claim No.
	--	
A	DE, A1, 3 716 318 (BTS) 24 November 1988 (24.11.88), see abstract; claims 1,2,7-9.	1-14

ANHANG | ANNEX | ANNEXE

zum internationalen Recherchenbericht über die internationale Patentanmeldung Nr.

to the International Search Report to the International Patent Application No.

au rapport de recherche international relatif à la demande de brevet international n°

PCT/EP 92/02230 SAE 64827

In diesem Anhang sind die Mitglieder der Patentfamilien der im obengenannten internationalen Recherchenbericht angeführten Patentdokumente angegeben. Diese Angaben dienen nur zur Unterrichtung und erfolgen ohne Gewähr.

This Annex lists the patent family members relating to the patent documents cited in the above-mentioned international search report. The Office is in no way liable for these particulars which are given merely for the purpose of information.

La présente annexe indique les membres de la famille de brevets relatifs aux documents de brevets cités dans le rapport de recherche international visée ci-dessus. Les reseignements fournis sont donnés à titre indicatif et n'engagent pas la responsibilité de l'Office.

Im Recherchenbericht angeführtes Patentdokument Patent document cited in search report Document de brevet cité dans le rapport de recherche	Datum der Veröffentlichung Publication date Date de publication	Mitglied(er) der Patentfamilie Patent family member(s) Membre(s) de la famille de brevets	Datum der Veröffentlichung Publication date Date de publication
US A 5079628	07-01-92	JP A2 2060252	28-02-90
		JP A2 1174124	10-07-89
EP A2 257947	02-03-88	CA A1 1270304	12-06-90
		EP A3 257947	02-05-90
		JP A2 63060643	16-03-88
		US A 4807222	21-02-89
US A 4665519	12-05-87	CA A1 1243730	25-10-88
GB A 2125257		AU A1 15474/83	09-02-84
		DK A0 3565/83	01-08-83
		DK A 3565/83	05-02-84
		EP A2 100594	15-02-84
		EP A3 100594	13-11-85
		GB A1 2125257	29-02-84
		GB B2 2125257	26-03-86
		JP A2 59045743	14-03-84
		NO A 832656	06-02-84
		ZA A 8304246	28-03-84
DE A1 3716318	24-11-88	DE C2 3716318	14-08-91

Method and apparatus for utilizing channel state information in a wireless communication system

Publication number:	**TW230525** (B)
Publication date:	2005-04-01
Inventor(s):	LING FUNYUN [US]; WALLACE MARK [US]; WALTON JAY R [US]; KETCHUM JOHN W [US]; HOWARD STEVEN J [US] +
Applicant(s):	QUALCOMM INC [US] +
Classification:	
- international:	*H04B7/04; H04B7/06; H04B7/08; H04J99/00; H04L1/00; H04L27/26;* H04B7/04; H04B7/08; H04J99/00; H04L1/00; H04L27/26; (IPC1-7): H04B7/02; H04L1/02
- European:	H04B7/04M1; H04B7/06C1F1C; H04B7/06C1F1Q; H04B7/08C4J2; H04B7/08S; H04L1/00A1M; H04L1/00A5; H04L1/00A9B; H04L25/02C1; H04L25/02C11A1; H04L25/02C11A5; H04L25/03B9
Application number:	TW20020105077 20020318
Priority number(s):	US20010816481 20010323

Abstract of **TW 230525 (B)**

Techniques for transmitting data from a transmitter unit to a receiver unit in a multiple-input multiple-output (MIMO) communication system. In one method, at the receiver unit, a number of signals are received via a number of receive antennas, with the received signal from each receive antenna comprising a combination of one or more signals transmitted from the transmitter unit. The received signals are processed to derive channel state information (CSI) indicative of characteristics of a number of transmission channels used for data transmission. The CSI is transmitted back to the transmitter unit. At the transmitter unit, the CSI from the receiver unit is received and data for transmission to the receiver unit is processed based on the received CSI.

I230525

申請日期	91 3 18
案號	91105077
類別	H04L 1/02, H04B 7/02

（以上各欄由本局填註）

A4
C4

發明~~新型~~ 專利說明書

一、發明~~新型~~名稱	中文	用以於無線通信系統中使用頻道狀態資訊之方法及裝置
	英文	"METHOD AND APPARATUS FOR UTILIZING CHANNEL STATE INFORMATION IN A WIRELESS COMMUNICATION SYSTEM"
二、發明~~創作~~人	姓名	1.林方永 FUYUN LING 2.馬克 瓦勒斯 MARK WALLACE
	國籍	均美國 U.S.A.
	住、居所	1.美國加州聖地牙哥市摩豪斯大道5775號 5775 MOREHOUSE DRIVE SAN DIEGO, CALIFORNIA 92131, U.S.A. 2.美國麻薩諸塞州貝佛市麥德巷4號 4 MADEL LANE, BEDFORD, MASSACHUSETTS 01730, U.S.A.
三、申請人	姓名（名稱）	美商奎康公司 QUALCOMM INCORPORATED
	國籍	美國 U.S.A.
	住、居所（事務所）	美國加州聖地牙哥市摩豪斯大道5775號 5775 MOREHOUSE DRIVE, SAN DIEGO, CALIFORNIA 92121-1714, U.S.A.
	代表人姓名	菲力普 R. 華德渥斯 PHILIP R. WADSWORTH

申請日期	
案　　號	
類　　別	

（以上各欄由本局填註）

A4
C4

發明/新型 專利說明書

一、發明/新型名稱	中　文	
	英　文	
二、發明/創作人	姓　名	3.傑伊 R. 瓦頓 JAY R. WALTON 4.約翰 W. 凱奇 JOHN W. KETCHUM 5.史帝芬 J. 郝華德 STEVEN J. HOWARD
	國　籍	均美國 U.S.A.
	住、居所	3.美國麻薩諸塞州西佛市里奇武德路7號 7 LEDGEWOOD DRIVE, WESTFORD, MASSACHUSETTS 01886, U.S.A. 4.美國麻薩諸塞州哈佛市甜莓巷37號 37 CANDLEBERRY LANE, HARVARD, MASSACHUSETTS 01451, U.S.A. 5.美國麻薩諸塞州亞西蘭市賀奇路75號 75 HERITAGE AVENUE, ASHLAND, MASSACHUSETTS 01721, U.S.A.
三、申請人	姓　名（名　稱）	
	國　籍	
	住、居所（事務所）	
	代表人姓　名	

I230525

（由本局填寫）

承辦人代碼：	
大　　類：	
ＩＰＣ分類：	

A6
B6

本案已向：

國（地區）　申請專利，申請日期：　　　案號：　　　，□有　□無主張優先權

美國　　2001年03月23日　　09/816,481　☑有　□無主張優先權

有關微生物已寄存於：　　　寄存日期：　　　，寄存號碼：

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訂

線

五、發明說明（ 1 ）

發明背景

範疇

本發明概括地說明資料通信；更特定言之，本發明說明一種用以利用(完全或部分)頻道狀態資訊改良一個無線通信系統性能之新穎和改良式方法及裝置。

背景

廣泛地部署無線通信系統 以提供各種不同的通信型式(像是聲音、資料...等等)。該等系統可以劃碼多向接近(CDMA)、劃時多向接近(TDMA)、直交分頻調變(OFDM)或其它的調變技藝為基礎。OFDM系統可使某些頻道環境具有高性能。

於一個陸地通信系統(例如一個蜂巢式系統、一個廣播系統、一個多頻道多點分配系統(MMDS)...等等)中，一個發送器單元中的一個射頻(RF)調變之信號可經由若干條傳輸路徑抵達一個接收器單元。該等傳輸路徑的特徵典型地基於若干因素(像是衰退和多路徑)而隨時間改變。

可利用多條傳輸和接收天線藉以"提供多樣性以防備不利的路徑效應"和"改良性能"。如該等傳輸和接收天線之間的傳輸路徑為線性獨立的(即未將一條路徑上的一個傳輸作成該等其它路徑上之傳輸的一個線性組合；一般來說，該等傳輸路徑就某個程度而言係線性獨立的)時，則正確接收一個傳輸之信號的可能性將隨著天線數的增加而提高。大體而言，將隨著傳輸和接收天線數的增加而增進多樣性及改良性能。

五、發明說明(2)

一個多輸入多輸出(MIMO)通信系統使用多條(N_T條)傳輸天線和多條(N_R條)接收天線傳輸資料。可將一個MIMO頻道分解成N_C個獨立的頻道，其中$N_C \le \min\{N_T, N_R\}$。亦將該等每一個獨立的頻道意指為該MIMO頻道的一個空間副頻道，其相對應至一個維數。如利用該等多條傳輸和接收天線所產生的附加dimensionality時，則該MIMO系統可改良性能。

因此，該技藝需要"利用頻道狀態資訊(CSI)取得一個MIMO系統所產生之附加dimensionality的優勢、進而改良系統性能"之方法。

發明概要

本發明的觀點係提供技藝用以"處理一個多輸入多輸出(MIMO)通信系統中收到的信號、以回復傳輸之信號"和"評估一個MIMO頻道的特徵"。可利用種種接收器處理計畫導出頻道狀態資訊(CSI)，以表示該等用於資料傳輸之傳輸頻道的特徵。接著，將該CSI回報給該發送器系統，及利用該CSI調整該信號處理(例如編碼、調變...等等)。以此方式，則可根據該等判定之頻道條件達成高性能。

本發明的一個特殊具體實施例提供一種用以於一個MIMO通信系通中將資料從一個發送器單元傳輸給一個接收器單元之方法。根據該方法，該接收器單元經由若干條接收天線接收若干個信號，其中該自每一條接收天線所收到的信號包括一個或多個傳輸自該發送器單元的信號組合。處理該等收到的信號(例如經由一個頻道關連矩陣反

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轉(CCMI)計畫、一個無偏差最小均方誤差(UMMSE)計畫、或其它的接收器處理計畫)以導出CSI，藉以表示若干用於資料傳輸之傳輸頻道的特徵。編碼該CSI，及將該CSI傳回給該發送器單元。該發送器單元接收該接收器單元中的CSI，及根據該收到的CSI處理傳輸給該接收器單元的資料。

該呈報之CSI可包含完全CSI或部分CSI。完全CSI包含所有成對傳輸/接收天線之間該傳播路徑其充分的全頻寬特徵描述(例如該可用頻寬上的振幅和相位)。部分CSI可包含例如該等傳輸頻道的信號/雜訊比(SNR)。可於該發送器單元上根據該每一個傳輸頻道的SNR評估編碼該每一個傳輸頻道的資料，及可根據一個依照該SNR評估所選取的調變計畫調變該每一個傳輸頻道的編碼資料。就完全CSI處理而言，亦在根據該收到的CSI執行傳輸之前、先預處理該等調變符號。

本發明尚提供執行本發明其種種、具體實施例和特性之方法、系統及裝置，將於下更詳細地說明。

圖示簡單說明

從下面提出的詳述、連同該等圖示將顯見本發明的特性、性質及優點，其中於該等所有的圖示中，同樣的參考記號相對應地視為同等，及其中：

圖1，為一個能夠執行本發明種種觀點和具體實施例之多輸入多輸出(MIMO)通信系統的圖示；

圖2A和2B，為一個能夠分別執行部分CSI處理和完全

五、發明說明（ 4 ）

CSI處理之MIMO發送器系統其一個具體實施例的方塊圖；

圖3，為一個利用直交分頻調變(OFDM)之MIMO發送器系統其一個具體實施例的方塊圖；

圖4，為一個能夠就不同傳輸型式提供不同的處理、且同樣使用OFDM之MIMO發送器系統其一部分的方塊圖；

圖5和圖6，為一個具有多條(N_R條)接收天線、且能夠分別根據一種頻道關連矩陣反轉(CCMI)技藝和一種無偏差最小均方誤差(UMMSE)處理一個資料傳輸之接收器系統其兩個具體實施例的方塊圖；

圖7A，說明該MIMO系統就三種接收器處理技藝和不同的SNR值之平均生產率；及

圖7B，說明該等三種根據該資料平面圖而產生之接收器處理技藝的累加機率分配函數(CDF)。

發明詳細說明

圖1為一個能夠執行本發明種種觀點和具體實施例之多輸入多輸出(MIMO)通信系統100的圖示。系統100包含一與一第二個系統150通信之第一個系統110。可運作系統100以利用一個天線、頻率及暫時多樣性(於下說明)的組合增加光譜效能、改良性能及增強彈性。一方面，可運作系統150以根據該呈報之CSI調整欲傳輸之資料的處理(例如編碼和調變)。

於系統110內，一個資料來源112將資料(即資訊位元)提供給一個傳輸(TX)資料處理器114，其中TX資料處理器114根據一個特殊的編碼計畫編碼該資料、根據一個特殊的交

五、發明說明（ 5 ）

錯計畫插入(即重新安排)該編碼後資料、及將該等插入的位元映射到一個或多個用以傳輸該資料之傳輸頻道的調變符號中。該編碼增加了該資料傳輸的可靠度。該交錯提供時間多樣性給該等編碼位元、准許根據該等用於該資料傳輸之傳輸頻道的一個平均信號/雜訊比(SNR)傳輸該資料、對付衰退、及更進一步移除用以形成每一個調變符號之編碼位元間的關連。當藉由多個頻率副頻道傳輸該等編碼位元時，則該交錯尚可提供頻率多樣性。根據本發明的一個觀點，係根據系統110可利用的完全CSI或部分CSI執行編碼、交錯及符號映射(或前述三者的一個組合)，如圖1中所示。

可根據許多計畫於發送器系統110上執行編碼、交錯及符號映射。於2001年2月1日提出之美國專利申請案序號09/776,073、定名為"一個無線通信系統之編碼計畫"中說明一種特殊的計畫，其中指定給本申請案的受讓人，及以引用的方式併入本文中。

MIMO系統100同時在該通信鏈結的傳輸端和接收端使用多條天線。可利用該等傳輸和接收天線提供不同的空間多樣性形式，包含傳輸多樣性和接收多樣性。空間多樣性的特徵為使用多條傳輸天線和一條或多條接收天線。傳輸多樣性的特徵為藉由多條傳輸天線傳輸資料。典型地說，當自該等傳輸天線傳輸該資料時，則執行附加處理、以達成該希望的多樣性。例如，可及時延遲或重新安排該傳輸自不同傳輸天線的資料、可藉由該等可利用的傳輸天線編碼

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五、發明說明（ 6 ）

和插入該傳輸自不同傳輸天線的資料...等等。接收多樣性的特徵為在多條接收天線上接收該等傳輸之信號，且僅藉著經由不同的信號路徑接收該等信號以達成接收多樣性。

可以若干種不同的通信狀態(模式)運作系統100，其中每一種通信狀態均使用天線、頻率或暫時多樣性、或前述三者的一個組合。該等通信狀態可包含例如一種"多樣性"通信狀態和一種"MIMO"通信狀態。該多樣性通信狀態使用多樣性改良該通信鏈結的可靠度。於該多樣性通信狀態的一個公用應用(亦意指為一種"純"多樣性通信狀態)中，係將資料從所有可利用的傳輸天線中將資料傳輸給一種領受的接收器系統。當該等資料傳送率需求低的時、或當該SNR低的時、或當該等兩者都低的時，則可利用該等純多樣性通信狀態。該MIMO通信狀態在該通信鏈結的兩端均使用天線多樣性(即多條傳輸天線和多條接收天線)，且通常利用該MIMO通信狀態改良可靠度該通信鏈結的容量。該MIMO通信狀態尚可使用頻率和/或暫時多樣性、連同該天線多樣性。

系統100尚可利用直交分頻調變(OFDM)，以有效地將該作業頻率波段分成若干個(L個)頻率副頻道(即頻率貯藏箱)。可於每一個時間磁格(即一段可依存該頻率副頻道頻寬的特殊時間區間)在該等每一個頻率副頻道上傳輸一個調變符號。

可運作系統100以經由若干個傳輸頻道傳輸資料。如上提及，可將一個MIMO頻道分解成N_C個獨立的頻道，其中

五、發明説明（ 7 ）

$N_C \leq \min\{N_T, N_R\}$。亦將該等每一個獨立的頻道意指為該MIMO頻道的一個空間副頻道。就一個未利用OFDM的MIMO系統而言，僅可有一個頻率副頻道，且可將每一個空間副頻道意指為一個"傳輸頻道"。就一個利用OFDM的MIMO系統而言，可將每一個頻率副頻道的每一個空間副頻道意指為一個傳輸頻道。又就一個未以該MIMO通信狀態運作的OFDM系統而言，僅有一個空間副頻道，且可將每一個頻率副頻道意指為一個傳輸頻道。

如利用該等多條傳輸和接收天線所產生的附加dimensionality時，則一個MIMO系統可改良性能。雖然此未必要求需了解該發送器上的CSI時，但當該發送器具有CSI(描寫從該等傳輸天線到該等接收天線的傳輸特徵)時，則可能增加系統的效率和性能。可將CSI類分為"完全CSI"亦或"部分CSI"。

完全CSI包含包含該$N_T x N_R$ MIMO矩陣中每一對傳輸-接收天線之間該傳播路徑其整個系統頻寬上充分的特徵描述(例如該振幅和相位)。完全CSI處理意味著：(1)可在該發送器和該接收器上利用的頻道特徵描述；(2)該發送器計算該MIMO頻道的特徵模式(於下説明)、判定欲在該等特徵模式上傳輸的調變符號、線性地事先制約(過濾)該等調變符號、及傳輸該等事先制約的調變符號；及(3)該接收器根據該頻道特徵描述執行該線性傳輸處理的一個互補處理(例如空間匹配過濾)，以計算每一個傳輸頻道(即每一種特徵模式)所需的該等N_C個空間匹配過濾係數。完全CSI處

五、發明說明（ 8 ）

理尚需根據該頻道的特徵值(於下說明)處理該每一個傳輸頻道的資料(例如選擇該等適當的編碼和調變計畫)，以導出該等調變符號。

部分CSI可包含例如該等傳輸頻道的信號/雜訊比(SNR)(即一個不具OFDM之MIMO系統其每一個空間副頻道的SNR；或一個具有OFDM之MIMO系統其每一個空間副頻道之每一個頻率副頻道的SNR)。部分CSI處理可意味著根據該頻道的SNR處理該每一個傳輸頻道的資料(例如選擇該等適當的編碼和調變計畫)。

參考圖1，一個TX MIMO處理器120接收和處理TX資料處理器114中的調變符號，以提供適合藉由該MIMO頻道傳輸的符號。TX MIMO處理器120所執行的處理係視使用完全CSI處理抑或部分CSI處理而定，將於下更詳細地說明。

就完全CSI處理而言，TX MIMO處理器120可將該等調變符號分工和事先制約。而就部分CSI處理而言，TX MIMO處理器120僅可將該等調變符號分工。將於下更詳細地說明該完全CSI MIMO處理和該部分CSI MIMO處理。就一個使用完全CSI處理、但未使用OFDM之MIMO系統而言，TX MIMO處理器120提供一個事先制約調變符號流給每一條傳輸天線，其中"一個事先制約調變符號/時間磁格"。每一個事先制約的調變符號為該等N_C個空間副頻道其於一個特定時間磁格上之N_C個調變符號的一個線性(和加權)組合。就一個使用完全CSI處理和OFDM之MIMO系統而言，TX MIMO處理器120提供一個事先制約調變符號向量流給

五、發明說明（ 9 ）

每一條傳輸天線，其中每一個向量包含一特定時間磁格上該等L個頻率副頻道其L個事先制約的調變符號。就一個使用部分CSI處理、但未使用OFDM之MIMO系統而言，TX MIMO處理器120提供一個調變符號流給每一條傳輸天線，其中"一個調變符號/時間磁格"。而就一個使用部分CSI處理和OFDM之MIMO系統而言，TX MIMO處理器120提供一個調變符號向量流給每一條傳輸天線，其中每一個向量包含一特定時間磁格上該等L個頻率副頻道的L個調變符號。就上述所有的事例而言，由一個個別的調變器(MOD) 122接收和調變每一個(無制約亦或事先制約的)調變符號流或每一個調變符號向量流，及經由一條相關的天線124傳輸該等每一個調變符號流或調變符號向量流。

於圖1中所示之具體實施例中，接收器系統150包含若干條接收天線152，其接收該等傳輸之信號和將該等收到的信號提供給個別的解調器(DEMOD) 154。每一個解調器154執行調變器122上所執行之處理的一個互補處理。將所有解調器154中該等解調變之符號提供給一個接收(RX) MIMO處理器156，及以一種下述之方式處理該等解調變之符號。繼之，將該等傳輸頻道中該等收到的調變符號提供給一個RX資料處理器158，其中RX資料處理器158執行TX資料處理器114其所執行之處理的一個互補處理。於一種特殊設計中，RX資料處理器158提供表示該等收到之調變符號的位元數值、deinterleave該等位元數值、及將該等deinterleave之數值解碼以產生解碼之位元、並接者將該等

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五、發明說明（ 10 ）

解碼之位元提供給一個資料槽160。將該收到的符號demap、deinterleave及解碼與在發送器系統110上將該符號映射、插入及編碼成係互補的。將於下更詳細地說明接收器系統150的處理。

一個MIMO系統的空間副頻道(或更概括言之，一個具有或不具OFDM之MIMO系統中的傳輸頻道)典型地經歷不同的鏈結條件(例如不同的衰退和多路徑效應)，且可達到不同的SNR。因此，該等每一個傳輸頻道的容量可為各自不同的。可由資訊位元傳送率(即每一個調變符號的資訊位元數)量化該容量，其中可在一個特殊的性能等級下、以該資訊位元傳送率在每一個傳輸頻道上傳輸。此外，該等鏈結條件典型地隨時間而變。從而該等傳輸頻道其支援之資訊位元傳送率亦隨時間而變。為了更完全地利用該等傳輸頻道的容量，故可判定(典型地在該接收器單元上判定)描寫該等鏈結條件的CSI和可將該CSI提供給該發送器單元，以便可相應地調整(或調適)該處理。本發明的觀點提供技藝判定CSI和利用(完全或部分)該CSI改良系統性能。

藉由部分CSI處理之MIMO發送器系統

圖2A為一個MIMO發送器系統110a其一個具體實施例之方塊圖，其為圖1中系統100其發送器部分的一個具體實施例。發送器系統110a(未利用OFDM)能夠根據接收器系統150所呈報的部分CSI調整其處理。系統110a包含：(1)一個接收和處理資訊位元、以提供調變符號之TX資料處理器114a；及(2)一個將該等N_T條傳輸天線的調變符號分工之

TX MIMO處理器120a。

TX資料處理器114a為圖1中TX資料處理器114的一個具體實施例，其中亦可將其它許多落在本發明範疇內的設計用在TX資料處理器114上。於圖2A中所示之特殊具體實施例中，TX資料處理器114a包含一個編碼器202、一個頻道交錯器204、一個截孔器206及一個符號映射元件208。編碼器202接收該等資訊位元，及根據一個特殊的編碼計畫編碼該等資訊位元、以提供編碼位元。頻道交錯器204根據一個特殊的交錯計畫插入該等編碼位元，以提供多樣性。截孔器206將該等零個或多個插入之編碼位元截孔，以提供該希望的編碼位元數。及符號映射元件208將該等截孔之編碼位元映射到一個或多個用以傳輸該資料之傳輸頻道的調變符號中。

可藉由該等處理過的資訊位元編碼和多工化導引資料(已知形態的資料)，雖然為了簡化而未於圖2A中說明此。可於該等用以傳輸該等資訊位元的所有或部分傳輸頻道中傳輸(例如以一種劃時多工之方式傳輸)該等處理過的導引資料。如此項技藝中已知，該接收器可利用該導引資料執行頻道評估，將於下更詳細地說明。

如圖2A中所示，可根據接收器系統150所呈報的部分CSI調整該編碼和調變。於一個具體實施例中，藉由利用一個固定的基碼(例如一個速率1/3的turbo碼)和調整該截孔達成適合的編碼，以達成該希望的編碼速率，就如該用以傳輸資料之傳殊頻道其SNR所支援的編碼速率。或者，可根據

五、發明說明（ 12 ）

該呈報之部分CSI(如進入區塊202中的虛線箭頭所示)利用不同的編碼計畫。例如，可藉由一個獨立編碼將該等每一個傳輸頻道編碼。藉由該編碼計畫，則可利用一個連續的"無效化/同等化和干擾取消"接收器處理計畫偵測和解碼該等資料流，以對該等傳輸之資料流導出一個更可靠的評估。P. W. Wolniansky等人於一篇定名為"V-釋放：一種用以於豐富分散化無線頻道上達成超高資料傳送率之結構"(會刊ISSSE-98、義大利比薩城)的論說中說明一種上述之接收器處理計畫，其中以引用的方式併入本文中。

就每一個傳輸頻道而言，可將符號映射元件208設計成非截孔編碼位元的群組集，以形成非二進位符號，及將該等非二進位符號映射到一個相對應至一選取給該傳輸頻道之特殊調變計畫(例如QPSK、M-PSK(相位移調變)、M-QAM(正交振幅調變)、或其它的計畫)上之信號星座中的點中。每一個映射點相對應至一個調變符號上。可傳輸給一個特殊性能等級(例如百分之一的封包錯誤率)之每一個調變符號的資訊位元總數係視該傳輸頻道的SNR而定。如是，可根據該呈報之部分CSI選擇每一個傳輸頻道的編碼計畫和調變計畫。亦可根據該呈報之部分CSI調整該頻道交錯(如進入區塊204中的虛線箭頭所示)。

表1列示種種可用於若干SNR範圍上的編碼速率和調變計畫組合。可利用任一種可能的編碼速率和調變計畫組合達成該每一個傳輸頻道其支援的位元傳送率。例如，可利用下面達成"一個資訊位元/符號"：(1)一個1/2的編碼速率

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五、發明說明（ 13 ）

與QPSK調變；(2)一個1/3的編碼速率與8-PSK調變；(3)一個1/4的編碼速率與6-QAM，或其它的編碼速率和調變計畫組合。於表1中，利用QPSK、16-QAM及64-QAM係作為該等列示之SNR範圍。亦可利用其它落在本發明範疇內的調變計畫 像是8-PSK、32-QAM、128-QAM... 等等。

表 1

SNR範圍	資訊位元數/符號	調變符號	編碼位元數/符號	編碼速率
1.5-4.4	1	QPSK	2	1/2
4.4-6.4	1.5	QPSK	2	3/4
6.4-8.35	2	16-QAM	4	1/2
8.35-10.4	2.5	16-QAM	4	5/8
10.4-12.3	3	16-QAM	4	3/4
12.3-14.15	3.5	64-QAM	6	7/12
14.15-15.55	4	64-QAM	6	2/3
15.55-17.35	4.5	64-QAM	6	3/4
> 17.35	5	64-QAM	6	5/6

將TX資料處理器114a中的調變符號提供給一個TX MIMO處理器120a，其中TX MIMO處理器120a為圖1中TX MIMO處理器120的一個具體實施例。於TX MIMO處理器120a內，一個分工器214將該等收到的調變符號分工成若干個(N_T個)調變符號流，其中利用每一條天線的一個流傳輸該等調變符號。將每一個調變符號流提供給一個個別的調變器122。每一個調變器122將該等調變符號轉換成一個類比信號，且更進一步將該信號放大、過濾、正交調變及

五、發明說明(14)

向上轉換，以產生一個適合藉由該無線鏈結傳輸之調變信號。

如該空間副頻道總數小於該可利用的傳輸天線總數時(即 $N_C < N_T$)，則可利用不同的計畫傳輸該資料。其中一種計畫係產生 N_C 個調變符號流，及於該等一部分可利用的傳輸天線(即 N_C 條傳輸天線)上傳輸該等產生之 N_C 個調變符號流。未利用該等剩餘的傳輸天線((N_T-N_C)條)傳輸該資料。另一種計畫係利用該等(N_T-N_C)條額外的傳輸天線其所提供的附加自由度改良該資料傳輸的可靠性。就該計畫而言，可將每一個資料流編碼、儘可能地插入及藉由多條傳輸天線傳輸。使用多條傳輸天線傳輸一個資料將增加多樣性和提高可靠性，以預防不利的路徑效應。

具完全CSI處理之MIMO發送器系統

圖2B為一個能夠根據接收器系統150所呈報之完全CSI處理資料之MIMO發送器系統110b(未利用OFDM)其一個具體實施例之方塊圖。一個TX資料處理器114將該等資訊位元編碼、插入及符號映射，以產生調變符號。可根據該接收器系統所呈報之該可利用的完全CSI調整該編碼和調變，及可如上就MIMO發送器系統110a所作的說明執行該編碼和調變。

於一個TX MIMO處理器120b內，一個頻道MIMO處理器212將該等收到的調變符號分工成若干個(N_C個)調變符號流，其中利用每一個空間副頻道的一個流(即特徵模式)傳輸該等調變符號。就完全CSI處理而言，頻道MIMO處理器

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五、發明說明（ 15 ）

212於每一個時間磁格事先制約該等N_C個調變符號，以產生N_T個制約之調變符號，如下：

$$\begin{bmatrix} x_1 \\ x_2 \\ M \\ x_{N_T} \end{bmatrix} = \begin{bmatrix} e_{11}, & e_{12}, & & e_{1N_c} \\ e_{21}, & e_{22}, & & e_{2N_c} \\ & & & \\ e_{N_T1}, & e_{N_T1}, & & e_{N_TN_c} \end{bmatrix} \cdot \begin{bmatrix} b_1 \\ b_2 \\ M \\ b_{N_C} \end{bmatrix} \qquad \text{方程式(1)}$$

其中b_1、b_2、...及b_{Nc}分別為該等空間副頻道1、2...、N_{Nc}的調變符號及其中可利用例如M-PSK、M-QAM或其它的調變計畫產生該等N_C個調變符號的每一個調變符號；

e_{ij}為一個與從該等傳輸天線到該等接收天線之傳輸特徵有關的特徵向量矩陣E的元件；及

x_1、x_2、...、x_{NT}為該等事先制約之調變符號，可表示成：

$$x_1 = b_1 \cdot e_{11} + b_2 \cdot e_{12} + \ \dots\ + b_{N_C} \cdot e_{1N_C}\ ,$$
$$x_2 = b_1 \cdot e_{21} + b_2 \cdot e_{22} + \ \dots\ + b_{N_C} \cdot e_{2N_C} \qquad \text{，及}$$
$$x_{N_T} = b_1 \cdot e_{N_T1} + b_2 \cdot e_{N_T2} + \ \dots\ + b_{N_C} \cdot e_{N_TN_C}\ .$$

可由該發送器計算該特徵向量矩陣E，或可由該接收器將該特徵向量矩陣E提供給該發送器。

就完全CSI處理而言，每一個給一條特殊傳輸天線的制約之調變符號x_i代表多達N_C個空間副頻道其(加權)調變符號的一個線性組合。該用於該每一個調變符號x_i上的調變計畫係以該特徵模式的有效SNR為基礎，且係與一個特徵值λ_i成比例(於下說明)。可使該等N_C個調變符號其每一個用以產生每一個事先制約之調變符號的調變符號與一個不

同的信號星座相關聯。就每一個時間磁格而言，一個分工器214將頻道MIMO處理器212所產生的該等N_T個事先制約之調變符號分工，並提供給N_T個調變器122。

可根據該等可利用的CSI和該等選取之傳輸天線執行該完全CSI處理。亦可選擇性和動態地致能和抑制該完全CSI處理。例如，可於一個特殊的資料傳輸時致能該完全CSI處理，且可於其它的資料傳輸時抑制該完全CSI處理。可於某些條件下致能該完全CSI處理，例如當該通信鏈結具有足夠的SNR時。

具OFDM之MIMO發送器系統

圖3為一個利用OFDM和能夠根據完全或部分CSI調整其處理之MIMO發送器系統110c其一個具體實施例之方塊圖。一個TX資料處理器114將該等資訊位元編碼、插入、截孔及符號映射，以產生調變符號。可根據該接收器系統所呈報之該可利用的完全或部分CSI調整該編碼和調變。就一個具有OFDM之MIMO系統而言，可於多個頻率副頻道上和可自多條傳輸天線傳輸該等調變符號。當以一種純MIMO通信狀態運作時，則該在每一個頻率副頻道上和自每一條傳輸天線的傳輸代表非備份資料。

於一個MIMO處理器120c內，一個分工器(DEMUX) 310接收該等調變符號，及將該等調變符號分工成若干個副頻道符號流$S_1 \sim S_L$。利用每一個頻率副頻道的一個副頻道符號流傳輸該等符號。

就完全CSI處理而言，繼之將每一個副頻道符號流提供

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給一個個別的副頻道MIMO處理器312。每一個副頻道MIMO處理器312將該等收到的副頻道符號流分工成若干個(多達N_C個)符號支流，其中利用每一個空間副頻道有一個符號支流傳輸該等調變符號。就一個OFDM系統中的完全CSI處理而言，根據"每一個頻率副頻道"導出和應用該等特徵模式。如是，每一個副頻道MIMO處理器312根據方程式(1)事先制約多達N_C個調變符號，以產生事先制約之調變符號。一個特殊頻率副頻道其一條特殊傳輸天線的每一個事先制約之調變符號代表多達N_C個空間副頻道其(加權)調變符號的一個線性組合。

就完全CSI處理而言，一個個別的分工器314將每一個副頻道MIMO處理器312其於每一個時間磁格所產生的(多達)N_T個事先制約之調變符號分工，並提供給(多達)N_T個符號組合器316a~316t。例如，指定給頻率副頻道1的副頻道MIMO處理器312a可提供多達N_T個事先制約之調變符號給天線1~N_T的頻率副頻道1。同樣地，指定給頻率副頻道L的副頻道MIMO處理器312l可提供多達N_T個符號給天線1~N_T的頻率副頻道L。

又就部分CSI處理而言，一個個別的分工器314將每一個副頻道符號流S分工，並提供給(多達)N_T個符號組合器316a~316t。就部分CSI處理而言，越過副頻道MIMO處理器312的處理。

每一個組合器316接收該等多達L個頻率副頻道的調變符號，將該等每一個時間磁格的符號組合至一個調變符號向

五、發明說明（ 18 ）

量V中，及將該調變符號向量提供給該下一個處理階段(即調變器122)。

如是，MIMO處理器120c接收和處理該等調變符號，以提供N_T個調變符號向量$V_1 \sim V_T$，其中每一條傳輸天線有一個調變符號向量。每一個調變符號向量V覆蓋一個單一的時間磁格，及使該調變符號向量V的每一個元件與一個具有一獨一無二、且於其上傳遞該調變符號之副載波之特殊頻率副頻道相關聯。如未以一種"純"MIMO通信狀態運作時，則該等某些調變符號向量可於不同傳輸天線的特殊頻率副頻道上擁有備份或冗位資訊。

圖3亦就OFDM說明調變器122的一個具體實施例。將MIMO處理器120c中的調變符號向量$V_1 \sim V_T$分別提供給調變器122a~122t。於圖3中所示之具體實施例中，每一個調變器122包含一個反轉快速傅利葉轉換(IFFT) 320、循環前置產生器322及一個向上轉換器324。

IFFT 320利用IFFT將每一個收到的調變符號向量轉換成其時間領域表示(意指為一個OFDM符號)。可將IFFT 320設計成對任意個頻率副頻道(例如8、16、32個...等等)執行該IFFT。於一個具體實施例中，就每一個轉換成一個OFDM符號之調變符號向量而言，循環前置產生器322重複該OFDM符號其時間領域表示的一部分，以形成一條特殊傳輸天線的一個傳輸符號。該循環前置確保該傳輸符號在多路徑延遲傳播面前保留其直交特性，藉以改良性能、以預防不利的路徑效應。於此不詳述此項技藝中已知之IFFT

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五、發明說明（ 19 ）

320和循環前置產生器322的執行。

接著，向上轉換324處理(例如轉換成一個類比信號、調變、放大及過濾)每一個循環前置產生器322中的時間領域表示(即該等每一條天線的傳輸符號)，以產生一個調變信號，並繼之經由個別的天線124傳輸該調變信號。

John A.C. Bingham於一篇定名為"資料傳輸之多載波調變：An Idea Whose Time Has Come"的論說(電子電機工程師協(IEEE)通信期刊，1990年5月)中更詳細地說明了OFDM調變，其中以引用的方式併入本文中。

一個通訊系統可傳輸若干不同的傳輸型式(例如聲音、打信號、資料、導引、...等等)。該等每一種傳輸可要求不同的處理。

圖4為一個能夠提供不同處理給不同傳輸型式、且亦使用OFDM之MIMO發送器系統110d其一部分之方塊圖。將該包含所有欲由系統110d傳輸之資訊位元的聚合輸入資料提供給一個分工器408。分工器408將該輸入資料分工成若干個(K個)頻道資料流B_1~B_K。每一個頻道資料流可相對應至例如一個打信號頻道、一個廣播頻道、一個語音呼叫、或一個封包資料傳輸上。將每一個頻道資料流提供給一個個別的TX資料處理器114，其中TX資料處理器114利用一個選取給頻道資料流的特殊編碼計畫編碼該資料、根據一個特殊的交錯計畫插入該等編碼後資料、及將該等插入的位元映射到一個或多個用以傳輸該頻道資料流之傳輸頻道的調變符號中。

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五、發明說明（ 20 ）

可根據"每一個傳輸"(即"每一個頻道資料流"，如圖4中所示)執行該編碼。然而，亦可編碼一組頻率副頻道、一組空間副頻道、一組頻率副頻道與空間副頻道、每一個頻率副頻道上的該聚合輸入資料(如圖1中所示)、若干頻道資料流、一個頻道資料流的一部分，及亦可編碼每一個調變符號、或其它的時間、空間和頻率單位。

可於一個或多個頻率副頻道上和經由每一個頻率副頻道的一個或多個空間副頻道傳輸每一個TX資料處理器114中的調變符號流。一個TX MIMO處理器120d自TX資料處理器114中接收該等調變符號流。TX MIMO處理器120d可依照該每一個調變符號流欲使用的通信狀態將該調變符號流分工成若干個副頻道符號流。於圖4中所示之具體實施例中，於一個頻率副頻道上傳輸調變符號流S_1，及於L個頻率副頻道上傳輸調變符號流S_K。一個個別的副頻道MIMO處理器412處理該每一個頻率副頻道的調變流，由分工器414分工該每一個調變流，及由組合器416組合該每一個調變流(例如以類似圖3中所述之方式組合)、以形成每一條傳輸天線的一個調變符號向量。

一般而言，一個發送器系統根據描寫頻道傳輸能力的資訊編碼和調變每一個傳輸頻道的資料。該資訊典型地係用上述完全CSI或部分CSI的形式。該接收器系統典型地判定該等用於資料傳輸之傳輸頻道的完全/部分CSI，及回報給該發送器系統。該發送器系統接著利用該資訊相應地調整該編碼和調變。該等於此所述之技藝可適用多個MIMO、

五、發明說明（ 21 ）

OFDM或其它任何能夠支援多個平行傳輸頻道的通信計畫(例如一個CDMA計畫)所支援的平行傳輸頻道。

於2000年3月22日提出之美國專利申請案序號09/532,492、定名為"利用多載波調變之高效率、高性能通信系統"中更詳細地說明MIMO處理，其中指定給本申請案的受讓人，及以引用的方式併入本文中。

MIMO接收器系統

本發明的觀點提供技藝"處理一個MIMO系統中該等收到的信號、以回復該傳輸之資料"和"評估該MIMO頻道的特徵"。接著，可將該等評估之頻道特徵回報給該發送器系統和用以調整該信號處理(例如編碼、調變...等等)。以此方式、根據該等判定之頻道條件達成高性能。該等於此所述之接收器處理技藝包含一種頻道關連矩陣反轉(CCMI)技藝、一種無偏差最小均方誤差(UMMSE)技藝及一種完全CSI技藝。將於下更詳細地說明上面所有的技藝。亦可利用其它落在本發明範疇內的接收器處理技藝。

圖1說明具有多條(N_R條)接收天線和能夠處理一個資料傳輸之接收器系統150。每一條接收天線152a~152r自多達N_T條傳輸天線中接收該等傳輸之信號，及將該等傳輸之信號的路線定訂至一個個別的解調器(DEMOD) 154(亦意指為一個前端處理器)上。例如，接收天線152a可自若干傳輸天線中接收若干傳輸之信號；同樣地，接收天線152r可接收多個傳輸之信號。每一個解調器154制約(例如過濾和放大)該收到的信號，將該制約之信號向下轉換成一個中頻

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或基頻，及將該向下轉換之信號數位化。每一個解調器154尚可藉由一個收到的導引解調變該等數位化樣品，以產生收到的調變符號 其中將該等收到的調變符號提供給RX MIMO處理器156。

如該資料傳輸係使用OFDM，則每一個解調器154將另外執行圖3中所示之調變器122其所執行之處理的一個互補處理。於該事例中，每一個解調器154包含一個產生該等樣品其轉換表示和提供一個調變符號向量流之FFT處理器(未顯示)，其中每一個向量包含L個頻率副頻道的L個調變符號。接著，將所有解調器其FFT處理器中的調變符號向量流提供給一個分工器/組合器(圖5中未顯示)，其中該分工器/組合器先將該每一個FFT處理器中的調變符號向量流"頻道化"成若干個(多達L個)副頻道符號流。可接著將該等每一個副頻道符號流提供給一個個別的RX MIMO處理器156。

就一個未利用OFDM之MIMO系統而言，可利用一個RX MIMO處理器156對該等N_R條收到之天線的調變符號執行該MIMO處理。又就一個利用OFDM之MIMO系統而言，可利用一個RX MIMO處理器156就該等L個用於資料傳輸之頻率副頻到的每一個頻率副頻道對該等N_R條收到之天線的調變符號執行該MIMO處理。

於一個具有N_T條傳輸天線和N_R條接收天線之MIMO系統中，可將該等N_R條接收天線其輸出端上收到的信號表示成：

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五、發明說明（ 23 ）

$\underline{r}=H\underline{x}+\underline{n}$ 方程式(2)

其中$\underline{r}$為該收到的符號向量(即自該MIMO頻道中輸出的N_R x 1向量，如於該等接受天線上測量的)；H為於一特殊時間上產生該頻道回應給該等N_T條傳輸天線和N_R條接收天線的N_R x N_T頻道係數矩陣；$\underline{x}$為該傳輸之符號向量(即輸入到該MIMO頻道中的N_T x 1向量)；及$\underline{n}$為一個代表noise plus interference之N_R x 1向量。該收到的符號向量$\underline{r}$包含於一特殊時間上經由N_R條接收天線所收到之N_R個信號的N_R個調變符號。同樣地，該傳輸之符號向量$\underline{x}$包含於一特殊時間上經由N_T條傳輸天線所傳輸之N_T個信號中的N_T個調變符號。

利用CCMI技藝之MIMO接收器

就該CCMI技藝而言，該接收器系首先對該收到的符號向量$\underline{r}$執行一個頻道匹配過濾作業，其中可將該過濾之輸出表示成：

$H^H\underline{r}=H^HH\underline{x}+H^H\underline{n}$ 方程式(3)

其中該註標"^{H}"表示調換和複合共軛。可利用一個平方矩陣R表示該頻道係數矩陣H與其共軛調換H^H的乘積(即$R=H^HH$)。

例如可從連同該資料一起傳輸的導引符號導出該頻道係數矩陣H。為了執行最佳化接收和評估該等傳輸頻道的SNR，通常將某些已知的符號插入該傳輸資料流中和於一個或多個傳輸頻道上傳輸該等已知的符號係合宜的。亦將該等已知的符號意指為導引符號或導引信號。可於若干此

五、發明說明(24)

項技藝可利用的論說中找到用以根據一個導引信號或該資料傳輸評估一個信號傳輸頻道之方法。F. Ling於一篇定名為"與應用參考-協助之一致CDMA通信之最佳化接收、性能界限及近路速率分析"(IEEE會報、1999年10月)中說明一種像這樣的頻道評估法。可將該頻道評估法或其它的頻道評估法擴增成矩陣形式,以導出該頻道係數矩陣H。

可以反轉R乘該信號向量$H^H\underline{r}$以獲得該傳輸之符號向量$\underline{x'}$的一個評估,其中可表示成:

$$\begin{aligned} x' &= R^{-1}H^H\underline{r} \\ &= \underline{x}+R^{-1}H^H\underline{n} \\ &= \underline{x}+\underline{n'} \end{aligned} \qquad \text{方程式(4)}$$

依照上面的方程式,將觀察到可藉由匹配過濾(即乘以該矩陣H^H)該等收到的符號向量$\underline{r}$和繼之以該反轉平方矩陣R^{-1}乘該過濾之結果、以回復該傳輸之符號向量$\underline{x}$。

可如下判定該等傳輸頻道的SNR首先根據該收到的信號計算該雜訊向量$\underline{n}$的自動關連矩陣ϕ_{nn}。一般而言,ϕ_{nn}為一個Hermitian矩陣,即其為複合共軛對稱的如該頻道雜訊的元件為無關連的時,且如該等元件更進一步為獨立和同一分配(iid)的時,則可將該雜訊向量$\underline{n}$的自動關連矩陣ϕ_{nn}表示成:

$$\phi_{nn}=\sigma_n^2 I\text{,及} \qquad \text{方程式(5)}$$

$$\phi_{nn}^{-1}=\frac{1}{\sigma_n^2}I\text{,}$$

其中I為該恒等矩陣(即沿著該矩陣的對角線均為一;該矩陣中的其它部分則為零值);及σ_n^2為該等收到之信號的雜

五、發明說明（ 25 ）

訊變異數。可將該後處理雜訊向量n'的自動關連矩陣$\phi_{n'n'}$(即於該匹配過濾和預先乘以該矩陣R^{-1}之後)表示成：

$$\phi_{n'n'}=E[\underline{n}'\underline{n}'^H]$$
$$=\sigma_n^2 R^{-1} \qquad \text{方程式(6)}$$

根據方程式(6)，該後處理雜訊n'其第i個元件的雜訊變異數σ_n^2等於$\sigma_n^2 \lambda_{ii}$，其中λ_{ii}為R^{-1}的第i個對角線元件。就一個未利用OFDM之MIMO系統而言，該第i個元件代表該第i條接收天線。如利用OFDM時，則可將該註標"i"分解成一個註標"jk"，其中"j"代表該第j個頻率副頻道，及"k"代表相對應至該第k條接收天線上的第k個空間副頻道。

就該CCMI技藝而言，可將該處理過之收到的符號向量其第i個元件(即x'的第i個元件)的SNR表示成：

$$SNR_i=\frac{\overline{|x'_i|^2}}{\sigma_{n'}^2} . \qquad \text{方程式(7)}$$

如該第i個傳輸之符號$\overline{|x'_i|^2}$的變異數平均等於一(1.0)時，則可將該接收符號向量的SNR表示成：

$$SNR_i=\frac{1}{\lambda_{ii}\sigma_n^2} .$$

可藉由$1/\sqrt{\lambda_{ii}}$量化該收到之符號向量的第i個元件，以常態化該雜訊變異數。

可將該等N_R條接收天線中的量化信號加總起來，以形成一個組合之信號，其中可表示成：

$$x'_{total}=\sum_{i=1}^{N_R}\frac{x'_i}{\lambda_{ii}} . \qquad \text{方程式(8)}$$

五、發明說明（ 26 ）

接著，該組合之信號的SNR"SNR_{total}"將有一個等於該等N_R條接收天線中其信號SNR總和之最大組合SNR。可將該組合之SNR表示成：

$$SNR_{total}=\sum_{i=1}^{N_R}SNR_i=\frac{1}{\sigma_n^2}\sum_{i=1}^{N_R}\frac{1}{r_{ii}}\text{。} \qquad \text{方程式(9)}$$

圖5說明一個RX MIMO處理器156a之一個具體實施例，其能夠執行該上述的CCMI處理。於RX MIMO處理器156a內，一個多工器512將該等N_R條接收天線中的調變符號多工化，以形成一個收到之調變符號向量(r)流。可根據類似於傳統導引協助之單一和多載波系統的導引信號評估該頻道係數矩陣H，如此項技藝中已知的。接著，根據如上所示之$R=H^HH$計算該矩陣R繼之，一個匹配過濾器514過濾該等收到的調變符號向量r，其中以該共軛調換頻道係數矩陣H^H預先乘每一個向量r，如上方程式(3)中所示。另外，一個乘法器516以該反轉平方矩陣R^{-1}預先乘該等過濾之向量，以形成該傳輸之調變符號向量x的一個評估x'，如上方程式(4)中所示。

就某些通信狀態而言，可將該等所有天線中用於傳輸該頻道資料流的副頻道符號流提供給一個組合器518，其中組合器518將時間、空間及頻率的冗位資訊組合起來。接著，將該等組合之調變符號x''提供給RX資料處理器158。就其它某些通信狀態而言，可將該等頻估之調變符號x'直接提供給RX資料處理器158(圖5中未顯示)。

如是，RX MIMO處理器156a產生若干個相對應至該發

五、發明說明（ 27 ）

送器系統上所利用之該等若干個傳輸頻道上的獨立符號流。每一個符號流包含後處理調變符號，其中該等後處理調變符號在該發送器系統的完全/部分CSI處理之前係相對應至該等調變符號上。接著，將該等(後處理)符號流提供給RX資料處理器158。

於RX資料處理器158內，將每一個後處理之調變符號的符號流提供給一個執行一解調變計畫(例如M-PSK、M-QAM)的個別解調變元件，其中該解調變計畫與該發送器系統用於處理該傳輸頻道的調變計畫互補。就該MIMO通信狀態而言，可繼之將該等所有指定之解調器中的解調變資料獨立地解碼；或可將該等所有指定之解調器中的解調變資料先多工化成一個頻道資料流 再繼之將該頻道資料流解碼，端視該發送器單元所使用的編碼和調變方法而定。可接著將每一個頻道資料流提供給一個執行一解碼計畫的個別解碼器，中該解碼計畫與該發送器單元用於該頻道資料流的計畫互補。該每一個解碼器中的解碼資料代表該頻道資料流其傳輸之資料的一個評估。

亦將該等評估之調變符號$\underline{x}'$和/或該等組合之調變符號$\underline{x}''$提供給一個CSI處理器520、以判定該等傳輸頻道的完全或部分CSI，及將該欲回報的完全/部分CSI提供給發送器系統110。例如，CSI處理器520可根據該收到的導引信號評估該第i個傳輸頻道的雜訊協方差矩陣ϕ_{nn}，及接著根據方程式(7)和(9)計算該SNR。類似於傳統導引協助之單一和多載波系統評估該SNR，如此項技藝中已知的。該等傳輸

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頻道的SNR包括回報給該發送器系統的部分CSI。另外，將該等調變符號提供給一個頻道評估器522和一個矩陣處理器524，以分別評估該頻道係數矩陣H和導出該平方矩陣R。一個控制器530耦合至RX MIMO處理器156a和RX資料處理器158上，及指導該等單元的作業。

利用UMMSE技藝之MIMO接收器

就該UMMSE技藝而言，該接收器系統以一個矩陣M乘該收到的符號向量$\underline{r}$，以導出該傳輸之符號向量$\underline{x}$的一個啟始MMSE評估$\hat{\underline{x}}$，其中可表示成：

$$\hat{\underline{x}}=M\underline{r} \qquad \text{方程式(10)}$$

選取該矩陣M，致使該啟始MMSE評估$\hat{\underline{x}}$和該傳輸之符號向量$\underline{x}$間之誤差向量$\underline{e}$的均方誤差最小(即$e=\hat{x}-x$)。

為了判定M，可先將一個成本函數ε表示成：

$$\begin{aligned}\varepsilon &= E\{\underline{e}^H\underline{e}\}\\ &=E\{[\underline{r}^H M^H-\underline{x}^H][M\underline{r}-\underline{x}]\}\\ &=E\{[\underline{r}^H M^H M\underline{r}-2\mathrm{Re}[\underline{x}^H M\underline{r}]+\underline{x}^H\underline{x}]\end{aligned}$$

為了使該成本函數ε最小化，可就M獲得該成本函數的一個導函數，及可將該結果設定成零，如下：

$$\frac{\partial}{\partial \mathbf{M}}\varepsilon = 2(\mathbf{HH}^H+\phi_{nn})\mathbf{M}^H-2\mathbf{H}=\mathbf{0}\ .$$

利用該等等式$E\{\underline{x}\underline{x}^H\}=I$、$E\{\underline{r}\underline{r}^H\}=HH^H+\phi_{nn}$及$E\{\underline{r}\underline{x}^H\}=H$獲得下面：

$$2(HH^H+\phi_{nn})M^H=2H$$

如是，可將該矩陣M表示成：

五、發明說明（ 29 ）

$$M=H^{H}(HH^{H}+\phi_{nn})^{-1}$$ 方程式(11)

可根據方程式(10)和(11)將該傳輸之符號向量$\hat{\underline{x}}$的啟始MMSE評估判定成：

$$\hat{\underline{x}}=M\underline{r}$$
$$=H^{H}(HH^{H}+\phi_{nn})^{-1}\underline{r}$$ 方程式(12)

可先根據特定x其平均在該加成雜訊上的平均值$\hat{\underline{x}}$判定該信號元件，以判定在該UMMSE技藝下該等傳輸頻道的SNR，其中可表示成：

$$\begin{aligned}E[\hat{\underline{x}}\mid\underline{x}] &= E[\mathbf{M}\underline{\mathbf{r}}\mid\underline{\mathbf{x}}]\\ &= \mathbf{H}^{H}(\mathbf{H}\mathbf{H}^{H}+\phi_{nn})^{-1}E[\underline{\mathbf{r}}]\\ &= \mathbf{H}^{H}(\mathbf{H}\mathbf{H}^{H}+\phi_{nn})^{-1}\mathbf{H}\underline{\mathbf{x}}\\ &= \mathbf{V}\underline{\mathbf{x}}\quad,\end{aligned}$$

其中將該矩陣V定義成：

$$V=\{v_{ij}\}$$
$$=MH$$
$$=H^{H}(HH^{H}+\phi_{nn})^{-1}H$$

利用該恒等

$$(\mathbf{H}\mathbf{H}^{H}+\phi_{nn})^{-1}=\phi_{nn}^{-1}-\phi_{nn}^{-1}\mathbf{H}(\mathbf{I}+\mathbf{H}^{H}\phi_{nn}^{-1}\mathbf{H})^{-1}\mathbf{H}^{H}\phi_{nn}^{-1}\ ,$$

可將該矩陣V表示成：

$$\mathbf{V}=\mathbf{H}^{H}\phi_{nn}^{-1}\mathbf{H}(\mathbf{I}+\mathbf{H}^{H}\phi_{nn}^{-1}\mathbf{H})^{-1}\ .$$

可將該啟始MMSE評估$\hat{\underline{x}}$，$\hat{x}_i$的第i個元件x_i表示成：

$$\hat{x}_i=v_{i1}x_1+\ldots+v_{ii}x_i+\ldots+v_{iN_R}x_{N_R}\ .$$ 方程式(13)

如$\hat{\underline{x}}$其所有的元件均無關連，且均具有零平均值時，則可將$\hat{\underline{x}}$其第i個元件的期望值表示成：

$$E[\hat{x}_i\mid\mathbf{x}]=v_{ii}x_i\ .$$ 方程式(14)

五、發明說明（ 30 ）

如方程式(14)中所示，$\hat{x}_i$為x_i的一個偏差評估。可根據該UMMSE技藝消除該偏差，以改良接收器的性能。可將x_i除以v_{ii}以獲得$\hat{x}_i$的一個無偏差評估。如是，可如下以一個對角線矩陣D_v^{-1}預先乘該偏差評估$\hat{\underline{x}}$，以獲得$\underline{x}$的無偏差最小均方誤差$\tilde{\underline{x}}$：

$$\tilde{\underline{\mathbf{x}}} = \mathbf{D}_{\mathbf{V}}^{-1}\hat{\underline{\mathbf{x}}}\ , \qquad \text{方程式(15)}$$

其中

$$\mathbf{D}_{\mathbf{V}}^{-1} = diag(1/v_{11}, 1/v_{22}, ..., 1/v_{N_R N_R})\ .$$

判定該noise plus interference，可將該無偏差評估$\tilde{\underline{x}}$該傳輸之符號向量$\underline{x}$間的誤差$\hat{\underline{e}}$表示成：

$$\begin{aligned}\hat{\underline{\mathbf{e}}} &= \underline{\mathbf{x}} - \mathbf{D}_{\mathbf{V}}^{-1}\hat{\underline{\mathbf{x}}} \\ &= \underline{\mathbf{x}} - \mathbf{D}_{\mathbf{V}}^{-1}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\underline{\mathbf{r}}\ .\end{aligned}$$

可將該誤差向量$\hat{\underline{e}}$的自動關連矩陣表示成：

$$\begin{aligned}\phi_{\hat{e}\hat{e}} &\equiv \mathbf{U} \equiv \{u_{ij}\} = E[\hat{\mathbf{e}}\hat{\mathbf{e}}^H] \\ &= \mathbf{I} - \mathbf{D}_{\mathbf{V}}^{-1}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\mathbf{H}(1 - \frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1}) - (1 - \frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1})\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\mathbf{H}\mathbf{D}_{\mathbf{V}}^{-1}.\end{aligned}$$

該誤差向量$\hat{\underline{e}}$其第i個元件的變異數等於u_{ii}。該誤差向量$\hat{\underline{e}}$的元件係相關連的。然而，可利用充分的交錯 致使能夠忽視該誤差向量$\hat{\underline{e}}$其元件間的關連性，及使得僅只該變異數才會影響系統性能。

如該頻道雜訊的元件為無關連和iid的，則可如方程式(5)中所示表示該頻道雜訊的關連矩陣。於該事例中，可將該誤差向量$\hat{\underline{e}}$的自動關連矩陣表示成：

$$\begin{aligned}\phi_{\hat{e}\hat{e}} &= \mathbf{I} - \mathbf{D}_{\mathbf{X}}^{-1}[\mathbf{I} - \sigma_n^2(\sigma_n^2\mathbf{I} + \mathbf{R})^{-1}](\mathbf{I} - \frac{1}{2}\mathbf{D}_{\mathbf{X}}^{-1}) - (\mathbf{I} - \frac{1}{2}\mathbf{D}_{\mathbf{X}}^{-1})[\mathbf{I} - \sigma_n^2(\sigma_n^2\mathbf{I} + \mathbf{R})^{-1}]\mathbf{D}_{\mathbf{X}}^{-1} \\ &= \mathbf{U} = \{u_{ij}\}.\end{aligned} \qquad \text{方程式(16)}$$

五、發明說明（ 31 ）

又如該頻道雜訊的元件為無關連的時，則

$$\mathbf{U}=\mathbf{I}-\mathbf{D}_{\mathbf{V}}^{-1}\mathbf{H}^{H}(\mathbf{H}\mathbf{H}^{H}+\phi_{nn})^{-1}\mathbf{H}(\mathbf{I}-\frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1})-(\mathbf{I}-\frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1})\mathbf{H}^{H}(\mathbf{H}\mathbf{H}^{H}+\phi_{nn})^{-1}\mathbf{H}\mathbf{D}_{\mathbf{V}}^{-1}. \quad \text{方程式(17)}$$

可將該相對應至該第i個傳輸之符號上的解調器輸出的SNR表示成：

$$SNR_i=\frac{E[\overline{|x_i|^2}]}{u_{ii}} \quad \text{方程式(18)}$$

如該等處理過的收到符號x_i之變異數$\overline{|x_i|^2}$平均等於一(1.0)時則可將該接收符號向量的SNR表示成：

$$SNR_i=\frac{1}{u_{ii}}.$$

圖6說明一個RX MIMO處理器156b之一個具體實施例，其能夠執行該上述的UMMSE處理。與該CCMI方法類似，可先根據該等收到的導引信號和/或資料傳輸評估該等矩陣H和ϕ_{nn}。接著，根據方程式(11)計算該加權係數矩陣M。於RX MIMO處理器156b內，一個多工器612乘該等N_R條接收天線中的調變符號，以形成一個收到之調變符號向量($\underline{r}$)流。繼之，一個乘法器614以該矩陣M預先乘該等收到的調變符號向量$\underline{r}$，以形成該傳輸之符號向量$\underline{x}$的一個評估$\hat{\underline{x}}$，如上方程式(10)中所示。另外，一個乘法器616以該對角線矩陣D_v^{-1}預先乘該評估$\hat{\underline{x}}$，以形成該傳輸之符號向量$\underline{x}$的一個無偏差評估$\tilde{\underline{x}}$，如上方程式(15)中所示。

又可依照該正在執行的特殊通信狀態，而可將該等所有天線中用於傳輸該頻道資料流的副頻道符號流提供給一個組合器618，其中組合器618將時間、空間及頻率的冗位資訊組合起來。接著，將該等組合之調變符號$\tilde{\underline{x}}''$提供給RX

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資料處理器158。又就其它某些通信狀態而言，可將該等頻估之調變符號$\tilde{\underline{x}}$直接提供給RX資料處理器158。

亦將該等無偏差評估之調變符號$\tilde{\underline{x}}$和/或該等組合之調變符號$\tilde{\underline{x}}''$提供給一個CSI處理器620，其中CSI處理器620判定該等傳輸頻道的完全或部分CSI，及將該欲回報的完全/部分CSI提供給發送器系統110。例如，CSI處理器620可根據方程式(16)~(18)評估該第i個傳輸頻道的SNR。該等傳輸頻道的SNR包括回報給該發送器系統的部分CSI。方程式(11)中計算出來的該最佳化M應已使該誤差向量的基準減到最小。根據方程式(16)計算D_V。

<u>利用完全CSI技藝之MIMO接收器</u>

就該完全CSI技藝而言，可如上方程式(2)中所示將該等N_R條接收天線其輸出端上收到的信號表示成：

$\underline{r}=H\underline{x}+\underline{n}$。

可將該頻道矩陣與其共軛調換的乘積所形成之該Hermitian矩陣的特徵向量分解表示成：

$H^H H = E\Lambda . E^H$，

其中E為該特徵向量矩陣；及Λ為一個特徵值對角線矩陣，其中E和Λ的維數均為$N_T x N_T$。該發送器事先利用該特徵向量矩陣E制約一組(N_T個)調變符號($\underline{b}$)，如上方程式(1)中所示。如是，可將該等N_T條傳輸天線中該等傳輸(事先制約)之調變符號表示成：

$\underline{x}=E\underline{b}$。

因$H^H H$為Hermitian，故該特徵向量矩陣為一元的。如是，

五、發明說明（ 33 ）

如該等元件$\underline{b}$具有相同的乘冪時，則該等元件$\underline{x}$亦具有相同的乘冪。接著，可將該收到的信號表示成：

$$\underline{r}=HE\underline{b}+\underline{n} \qquad \text{方程式}(19)$$

該接收器執行一個頻道匹配過濾作業，並繼之乘上該等右特徵向量。該等頻道匹配過濾和乘法作業的結果為一個向量$\underline{z}$，其中可表示成：

$$\underline{z}=E^H H^H HE\underline{b}+E^H H^H \underline{n}=\Lambda\,\underline{b}+\underline{n}' \qquad \text{方程式}(20)$$

其中可將該新雜訊項的協方差表示成：

$$E(\underline{\hat{n}\hat{n}}^H)=E(E^H H^H \underline{nn}^H HE)=E^H H^H HE=\Lambda \qquad \text{方程式}(21)$$

即該等雜訊元件係與該等特徵值所產生的變異數無關。$\underline{z}$的第i個元件的SNR為λ_i，即Λ的第i個對角線元件。

於前述美國專利申請案序號09/532,492中更詳細地說明完全CSI處理。

亦可利用圖5中所示之接收器具體實施例執行該完全CSI技藝。匹配過濾器514過濾該等收到的調變符號向量$\underline{r}$，其中以該共軛調換頻道係數矩陣H^H預先乘每一個向量$\underline{r}$，如上方程式(20)中所示。另外，乘法器516以該等右特徵向量E^H預先乘該等過濾之向量，以形成該調變符號向量$\underline{b}$的一個評估$\underline{z}$，如上方程式(20)中所示。就該完全CSI技藝而言，將矩陣處理器524架構成提供該等右特徵向量E^H。可如上述達成該隨後的處理(例如藉由組合器518和RX資料處理器158)。

就該完全CSI技藝而言，該發送器單元可根據該特徵值所產生的SNR選擇一個編碼計畫和一個調變計畫(即一個

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五、發明說明（ 34 ）

信號星座)給該等每一個特徵向量。倘若該等頻道條件未明顯地變更該接收器測量該CSI與該發送器呈報和用以事先制約該傳輸間的時間區間時，則該通信系統的性能可能相當於一組具已知SNR之獨立AWGN頻道的性能。

將完全或部分CSI回報給該發送器系統

可利用該此中所述之部分CSI(例如CCMI或UMMSE)亦或完全CSI技藝獲得該等收到之信號的其每一個傳輸頻道的的SNR。可接著經由一個反向頻道將該等傳輸頻道其判定之SNR回報給該發送器系統。藉由回饋該等傳輸頻道其傳輸之調變符號的SNR值(即每一個空間副頻道；且如使用OFDM時、則可能為每一個頻率副頻道)，則執行調適處理(例如調適編碼和調變)以改良該MIMO頻道的使用係可能的。就該等部分CSI回饋技藝而言，可於沒有完全的CSI下達成調適處理。就該等完全CSI回饋技藝而言，可將充分的資訊(不一定是該明確的特徵值和特徵模式)回饋給該發送器，以幫助計算該等每一個使用之頻率副頻道的特徵值和特徵模式。

就該CCMI技藝而言，將該等收到之調變符號的SNR值(例如於該第i個傳輸頻道上收到之符號的 $SNR_i = \overline{|x'_i|^2} / \sigma_n^2$ 或 $SNR_i = 1/\sigma_n^2 r_{ii}$)回饋給該發送器。就該UMMSE技藝而言，將該等收到之調變符號(例如於該第i個傳輸頻道上收到之符號的 $SNR_i = E[|x_i|^2]/u_{ii}$ 或 $SNR_i = 1/u_{ii}$，其中如上方程式(16)和(17)中所示計算 u_{ii})的SNR值回饋給該發送器。又就該完全CSI技藝而言，可將該等收到之調變符號的SNR值(例如於

五、發明說明(35)

該第i個傳輸頻道上收到之符號的$SNRi=\overline{|z_i|^2}/\sigma_n^2$或$SNR_i=\lambda_{ii}/\sigma_n^2$，其中$\lambda_{ii}$為該平方矩陣的特徵值)回饋給該發送器。就該完全CSI技藝而言，可更進一步判定該等特徵模式E，及將該等E回饋給該發送器。就該等部分和完全CSI技藝而言，該發送器系統利用該SNR調整該資料處理。又就該完全CSI技藝而言，其更進一步在傳輸該等調變符號之前、先利用該等特徵模式E事先制約該等調變符號。

可將該欲回報給該發送器的CSI全部、差別、或全部/差別地傳送。於一個具體實施例中，定期地呈報完全或部分CSI，及根據該先前傳輸之CSI傳送差別更新。如完全CSI的一個實例，該等更新可為該等呈報之特徵模式的更正(根據一個誤差信號)。典型地說，該等特徵值不像該等特徵模式一樣迅速地變更，故可以一個較低的速率更新該等特徵值。於另一個具體實施例中，僅當產生變更時(例如當該變更超過一個特殊閥時)才傳送該CSI，此可降低該回饋頻道的有效速率。如部分CSI的一個實例，僅將該等SNR變更時、才將該等SNR送回(例如差別地)。就一個OFDM系統(具有或不具MIMO)而言，可利用該頻率領域中的關連性准許減少該欲回饋的CSI量。如一個利用部分CSI之OFDM系統的一個實例，如該相對應至M個頻率副頻道其一個特殊空間副頻道上的SNR相同時，則可呈報該SNR、及當該條件為真的時之該第一個和該最後一個頻率副頻道。亦可利用其它用以減少該欲回饋給CSI之資料量、且落在本發明範疇內之壓縮和回饋頻道錯誤回復技

五、發明說明（ 36 ）

藝。

往回參考圖1，將RX MIMO處理器156所判定的完全或部分CSI(例如頻道SNR)提供給一個TX資料處理器162，其中TX資料處理器162處理該CSI，及將處理過的資料提供給一個或多個調變器154。調變器154更進一步制約該處理過的資料，及經由一個反向頻道將該CSI傳回給發送器系統110。

於系統110上，天線124接收該傳輸之回饋信號 調變器122將該傳輸之回饋信號解調變，及將其提供給一個RX資料處理器132。RX資料處理器132執行TX資料處理器162所執行之處理的一個互補處理，及回復該呈報之完全/部分CSI，繼之，將該回復之完全/部分CSI提供給TX資料處理器114，及TX MIMO處理器120利用該回復之完全/部分CSI調整該處理。

發送器系統110可根據接收器系統150中的完全/部分CSI(例如SNR資訊)調整(即調適)其處理。例如，可調整該每一個傳輸頻道的編碼、使得該資訊位元傳送率與該頻道SNR所支援的傳輸能力匹配。此外，可根據該頻道SNR選擇該傳輸頻道的調變計畫。亦可於本發明的範疇內調整其它的處理(例如交錯)。根據判定給該頻道的SNR調整每一個傳輸頻道的處理容許該MIMO系統達成高性能(即一種特殊性能等級的高生產率或高位元傳送率)。可將該調適的處理應用到一個單載波MIMO系統或一個植基於多載波之MIMO系統(例如一個利用OFDM之MIMO系統)上。

五、發明說明（ 37 ）

可根據許多種技藝調整該發送器系統的編碼和其調變計畫的選擇，於前述的美國專利申請案序號09/776,073中說明其中一種技藝。

該等部分(例如CCMI和UMMSE)和完全CSI技藝為接收器處理技藝，其容許一個MIMO系統利用因使用多條傳輸和接收天線而產生的附加dimensionality，此為使用MIMO的一項主要優點。該等CCMI和UMMSE技藝可容許在每一個時間磁格傳輸該相同的調變符號總數，就如一個利用完全CSI之MIMO系統一樣。然而，亦可將其它落在本發明範疇內之接收器處理技藝、連同該等此中所述之完全/部分CSI回饋技藝一起使用。相仿地，圖5和6代表一個能夠處理一MIMO傳輸、判定該等傳輸頻道特徵(即該SNR)、及將完全或部分CSI回報給該發送器系統之接收器系統的兩個具體實施例。可慎思其它落在本發明範疇內之植基於該等此中提出之技藝的設計和接收器處理技藝。

當僅回饋該全部收到的信號SNR或根據該SNR評估之可達到的全面生產率時，則該發送器系統亦可在不調適處理的情況下以一種筆直的方式利用該部分CSI技藝(例如CCMI和UMMSE技藝)。於執行時，根據該收到的SNR評估或該評估之生產率判定一個調變格式，且所有的傳輸頻道均使用該相同的調變格式。該方法可減少全面的系統生產率，且亦可大大地減少藉由該反向鏈結所送回的資訊量。

可藉由使用本發明的完全/部分回饋技藝改良系統性能。可計算該具部分CSI回饋之系統生產率，及比較該系

五、發明說明（ 38 ）

統生產率與該具完全CSI回饋的生產率。可將該系統生產率定義成：

$$C=\sum_{i=1}^{N_C}\log_2(1+\gamma_i)\ ,$$

其中γ_i為部分CSI技藝其每一個收到之調變符號的SNR；或為該完全CSI技藝其每一個傳輸頻道的SNR。可將不同處理技藝的SNR概述如下：

$$\gamma_i=\frac{1}{\sigma_n^2 r_{ii}}\ ,\qquad \text{for the CCMI technique}$$

$$\gamma_i=\frac{1}{u_{ii}}\ ,\qquad \text{for the UMMSE technique, and}$$

$$\gamma_i=\frac{\lambda_{ii}}{\sigma_n^2}\ ,\qquad \text{for full CSI technique.}$$

圖7A和7B說明一個使用部分CSI和完全CSI回饋技藝之4x4 MIMO系統的性能。從電腦模擬獲得該等結果。於該模擬中，將每一個頻道係數矩陣H的元件模仿成"平均值=0"和"變異數=1"的獨立高斯隨機變數。就每一個計算而言，產生若干個隨機矩陣realization，及將該等realization其算出之生產率平均以產生該平均生產率。

圖7A說明該MIMO系統其在不同SNR值之完全CSI、部分CSI CCMI及部分CSI UMMSE技藝下的平均生產率。可從圖7A中察知，該部分CSI UMMS技藝的生產率於高SNR值時約為該完全CSI生產率的75%；且於低SNR值時近似該完全CSI的生產率。該部分CSI CCMI技藝的生產率於高SNR值時約為該部分CSI UMMSE技藝之生產率的75%～90%；且於低SNR值時約比該UMMSE的生產率少30%。

五、發明說明（ 39 ）

圖7B說明根據該資料平面圖而產生之該等三種技藝的累加機率分配函數(CDF)。圖7B顯示就該CCMI技藝而言，當一個平均SNR=16分貝(dB)/傳輸頻道時 則約有5%的情況其生產率會小於"2每秒位元/赫茲"。另一方面，於該相同的SNR下，該UMMSE技藝的生產率在所有的情況下均大於"7.5每秒位元/赫茲"。如是，該UMMSE技藝其運行中斷的機率可能比該CCMI技藝低。

可藉由一個或多個數位信號處理器(DSP)、特別應用積體電路(ASIC)、處理器、微處理器、控制器、微控制器、欄位可程式行列閘(FPGA)、可程式邏輯裝置、其它的電子設備、或以上任何組合執行該等發送器和接收器系統的元件。亦可藉由一個處理器上執行的軟體執行該等此中所述之某些功能和處理。

可藉由一個軟體和硬體的組合執行本發明的觀點。例如可根據一個處理器(圖5中的控制器530和圖6中的控制器650)上執行的程式碼計算該等CCMI和UMMSE技藝的符號評估和導出該頻道SNR。

提供先前對該等揭示之具體實施例的所作的說明，以致能熟諳此藝者實行或利用本發明。熟諳此藝者將即刻顯見到可對該等具體實施例作種種修正，且可在未脫離本發明的精髓或範疇下將此中定義的通用原則應用到其它的具體實施例上。如是，並不希望將本發明限制在此中所示之具體實施例上，而是希望本發明符合此中揭示之原則和新穎特性的最寬廣範疇。

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B5

四、中文發明摘要（發明之名稱：用以於無線通信系統中使用頻道狀態資訊之方法及裝置）

用以於一個多輸入多輸出(MIMO)通信系通中將資料從一個發送器單元傳輸給一個接收器單元之技藝。其中一種方法該接收器單元經由若干條接收天線接收若干個信號，其中該自每一條接收天線所收到的信號包括一個或多個傳輸自該發送器單元的信號組合。處理該等收到的信號、以導出頻道狀態資訊(CSI)，藉以表示若干用於資料傳輸之傳輸頻道的特徵。將該CSI傳回給該發送器單元。該發送器單元接收該接收器單元中的CSI，及根據該收到的CSI處理傳輸給該接收器單元的資料。

英文發明摘要（發明之名稱："METHOD AND APPARATUS FOR UTILIZING CHANNEL STATE INFORMATION IN A WIRELESS COMMUNICATION SYSTEM"）

Techniques for transmitting data from a transmitter unit to a receiver unit in a multiple-input multiple-output (MIMO) communication system. In one method, at the receiver unit, a number of signals are received via a number of receive antennas, with the received signal from each receive antenna comprising a combination of one or more signals transmitted from the transmitter unit. The received signals are processed to derive channel state information (CSI) indicative of characteristics of a number of transmission channels used for data transmission. The CSI is transmitted back to the transmitter unit. At the transmitter unit, the CSI from the receiver unit is received and data for transmission to the receiver unit is processed based on the received CSI.

六、申請專利範圍

1. 一種用以於一個多輸入多輸出(MIMO)通信系通中將資料從一個發送器單元傳輸給一個接收器單元之方法，包括：

   於該接收器單元方：

   經由多條接收天線接收多個信號，其中該自每一條接收天線所收到的信號包括一個或多個傳輸自該發送器單元的信號組合；

   處理該等收到的信號、以導出頻道狀態資訊(CSI)，藉以表示多個用於資料傳輸之傳輸頻道的特徵；及

   將該CSI傳回給該發送器單元；及

   於該發送器單元方：

   接收該接收器單元中的CSI；及

   根據該收到的CSI處理傳輸給該接收器單元的資料。

2. 如申請專利範圍第1項之方法，其中該呈報之CSI包括該等每一個傳輸頻道的信號/雜訊比(SNR)評估。

3. 如申請專利範圍第2項之方法，其中該發送器單元上的處理包含：

   根據每一個傳輸頻道的SNR評估編碼該每一個傳輸頻道的資料。

4. 如申請專利範圍第3項之方法，其中係根據每一個傳輸頻道的SNR評估獨立地編碼該每一個傳輸頻道的資料。

5. 如申請專利範圍第3項之方法，其中該編碼包含：

   以一固定的基碼編碼該傳輸頻道的資料；及

   根據該傳輸頻道的SNR評估調整編碼位元的截孔。

六、申請專利範圍

6. 如申請專利範圍第3項之方法，其中該發送器單元上的處理尚包含：

按照一根據該傳輸頻道其SNR評估而選取之調變計畫調變每一個傳輸頻道的編碼資料。

7. 如申請專利範圍第1項之方法，其中該呈報之CSI包括該等多個傳輸頻道的特徵描述。

8. 如申請專利範圍第1項之方法，其中該呈報之CSI指示該等多個傳輸頻道的特徵模式和特徵值。

9. 如申請專利範圍第8項之方法，其中該發送器單元上的處理包含：

根據該等特徵值編碼該等傳輸頻道的資料。

10. 如申請專利範圍第9項之方法，其中獨立地編碼該每一個傳輸頻道的資料。

11. 如申請專利範圍第9項之方法，其中該發送器單元上的處理尚包含：

按照根據該等特徵值而選取之調變計畫調變該等傳輸頻道的編碼資料，以提供調變符號。

12. 如申請專利範圍第11項之方法，其中該發送器單元上的處理尚包含：

於傳輸該等調變符號之前、先根據該等特徵模式事先制約該等調變符號。

13. 如申請專利範圍第1項之方法，其中自該接收器單元中傳輸該全部的CSI。

14. 如申請專利範圍第13項之方法，其中定期地自該接收器

六、申請專利範圍

單元中傳輸該全部的CSI，及其中於完全傳輸之間傳輸該等CSI更新。

15. 如申請專利範圍第1項之方法，其中當偵測到該等頻道特徵變更超過一個特殊閥時，則傳輸該CSI。

16. 如申請專利範圍第8項之方法，其中以不同的更新速率傳輸該指示該等特徵模式和特徵值之CSI。

17. 如申請專利範圍第1項之方法，其中該接收器單元根據一個關連矩陣反轉(CCMI)處理導出該CSI。

18. 如申請專利範圍第17項之方法，其中該接收器單元上的CCMI處理包含：

處理該等收到的信號，以導出收到之調變符號；

根據一第一個矩陣過濾該等收到的調變符號，以提供過濾之調變符號，其中該第一個矩陣表示用於該資料傳輸之多條傳輸天線與該等多條接收天線之間的一頻道特徵評估；

以一第二個矩陣乘該等過濾之調變符號，以評估傳輸之調變符號；及

評估多個用於該資料傳輸之傳輸頻道的特徵。

19. 如申請專利範圍第18項之方法，尚包括：

根據一特殊的解調變計畫將該等調變符號評估解調變，以提供解調變符號。

20. 如申請專利範圍第19項之方法，尚包括：

根據一特殊的解碼計畫將該等解調變符號解碼。

21. 如申請專利範圍第18項之方法，尚包括：

六、申請專利範圍

組合冗位傳輸的調變符號評估，以提供組合之調變符號評估。

22. 如申請專利範圍第18項之方法，尚包括：

根據該等收到的調變符號導出一頻道係數矩陣；及

其中根據該頻道係數矩陣導出該第一個矩陣。

23. 如申請專利範圍第22項之方法，其中根據相對應至導引資料上之收到的調變符號導出該頻道係數矩陣。

24. 如申請專利範圍第18項之方法，其中該第二個矩陣為一個根據該第一個矩陣所導出之反轉平方矩陣。

25. 如申請專利範圍第1項之方法，其中該接收器單元根據一個無偏差最小均方誤差(UMMSE)處理導出該CSI。

26. 如申請專利範圍第25項之方法，其中該UMMSE處理包含：

處理該等收到的信號，以導出收到之調變符號；

以一第一個矩陣M乘該等收到的調變符號，以評估傳輸之調變符號；

根據該收到的調變符號評估多個用於該資料傳輸之傳輸頻道的特徵；及

其中選擇該第一個矩陣M，使得該等調變符號評估與傳輸之調變符號間的一均方誤差減到最小。

27. 如申請專利範圍第26項之方法，尚包括：

以一第二個矩陣乘該等調變符號評估，以無偏差地評估該等傳輸之調變符號；及

其中根據該等無偏差之調變符號評估評估該等傳輸頻

六、申請專利範圍

道的特徵。

28. 如申請專利範圍第27項之方法，尚包括：

根據該等無偏差之調變符號評估導出該第一個矩陣M，及使該等無偏差之調變符號評估與該等傳輸之調變符號間的該均方誤差減到最小。

29. 如申請專利範圍第1項之方法，其中該MIMO系統執行直交分頻調變(OFDM)。

30. 如申請專利範圍第29項之方法，其中就多個頻率副頻道的每一個頻率副頻道執行該收發器單元和該發送器單元上的處理。

31. 一種用以於一個多輸入多輸出(MIMO)通信系通中將資料從一個發送器單元傳輸給一個接收器單元之方法，包括：

於該接收器單元方：

經由多條接收天線接收多個信號，其中該自每一條接收天線所收到的信號包括一個或多個傳輸自該發送器單元的信號組合；

處理該等多個收到的信號，以評估傳輸自該發送器單元中之調變符號；

評估多個用於資料傳輸之傳輸頻道的信號/雜訊比(SNR)；及

將該等傳輸頻道的SNR評估傳回給該發送器單元；及

該發送器單元根據該等收到的SNR評估處理傳輸給該接收器單元的資料。

六、申請專利範圍

32. 如申請專利範圍第31項之方法，其中評估該等每一個傳輸頻道的SNR，及將該等每一個傳輸頻道的SNR評估傳回給該發送器單元。

33. 如申請專利範圍第31項之方法，尚包括：

於該接收器單元方：

導出該等多個用於資料傳輸之傳殊頻道的特徵描述；及

將該等特徵描述傳回給該發送器單元。

34. 如申請專利範圍第33項之方法，尚包括：

該發送器單元在將調變符號傳輸給該接收器單元之前、先根據該等多個傳輸頻道的特徵描述事先制約該等調變符號。

35. 如申請專利範圍第31項之方法，其中根據一個頻道關連矩陣反轉(CCMI)計畫處理該等收到的調變符號。

36. 如申請專利範圍第31項之方法，其中根據一個最小無偏差均方誤差(UMMSE)計畫處理該等收到的調變符號。

37. 如申請專利範圍第31項之方法，其中該發送器單元上的處理包含：

根據該收到之傳輸頻道的SNR評估編碼每一個傳輸頻道的資料。

38. 如申請專利範圍第37項之方法，其中該發送器單元上的處理尚包含：

按照一根據該收到之傳輸頻道的SNR評估而選取之調變計畫調變每一個傳輸頻道的編碼資料。

39. 一種多輸入多輸出(MIMO)系統，包括：

六、申請專利範圍

一接收器單元，包括：

多個前端處理器，其中架構成經由多條接收天線接收多個信號，及處理該等收到的信號、以提供收到之調變符號；

至少一耦合至該等前端處理器上之接收MIMO處理器，其中架構成接收和處理該等收到的調變符號，以導出頻道狀態資訊(CSI)，藉以表示多個用於資料傳輸之傳輸頻道的特徵；及

一有效耦合至該接收MIMO處理器上之傳輸資料處理器，其中架構成處理傳回給該發送器單元的CSI；及

一發送器單元，包括：

至少一個解調器，其中架構成接收和處理該等接收器單元中的一個或多個信號，以回復該傳輸之CSI；及

一傳輸資料處理器，其中架構成根據該回復之CSI處理傳輸給該接收器單元之資料。

40. 一種於一多輸入多輸出(MIMO)通信系統中之接收器單元，包括：

多個前端處理器，其中架構成經由多條接收天線接收多個傳輸之信號，及處理該等收到的信號、以提供收到之調變符號；

一有效耦合至該等多個前端處理器上之過濾器，其中架構成根據一第一個矩陣過濾該等收到的調變符號，以提供過濾之調變符號，及其中該第一個矩陣表示用於該資料傳輸之多條傳輸天線與該等多條接收天線之間的一

六、申請專利範圍

頻道特徵評估；

一耦合至該過濾器上之乘法器，其中架構成以一第二個矩陣乘該等過濾之調變符號，以評估傳輸之調變符號；

一耦合至該乘法器上之頻道品質評估器，其中架構成評估多個用於該資料傳輸之傳輸頻道的特徵，及提供頻道狀態資訊藉(CSI)、藉以表示該等評估之頻道特徵；及

一傳輸資料處理器，其中架構成接收和處理傳輸自該接收器單元中的CSI。

41. 如申請專利範圍第40項之接收器單元，尚包括：

一第二個評估器，其中架構成根據該等調變符號評估導出一頻道係數矩陣，及其中根據該頻道係數矩陣導出該第一個矩陣。

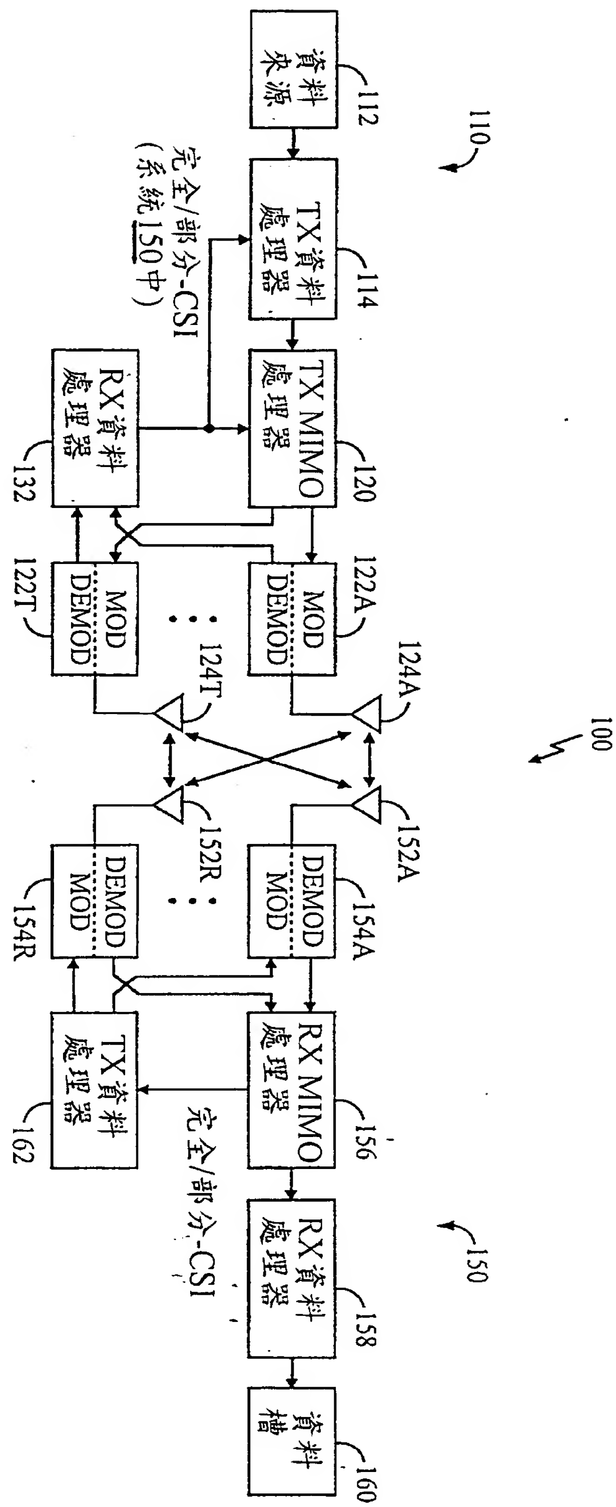
42. 如申請專利範圍第40項之接收器單元，其中該等傳輸頻道特徵評估包括信號/雜訊比(SNR)評估。

43. 如申請專利範圍第40項之接收器單元，尚包括：

一個或多個解調變元件，其中將每一個解調變元件架構成根據一特殊的解調變計畫接收和解調變一個別之調變符號評估流，以提供一解調變符號流。

44. 如申請專利範圍第43項之接收器單元，尚包括：

一個或多個解碼器，其中將每一個解碼器架構成根據一特殊的解碼計畫接收和解碼一解調變符號流，以提供解碼資料。



100
110
112
資料來源
114
TX資料處理器
120
TX MIMO處理器
完全/部分-CSI
(系統150中)
132
RX資料處理器
122A
MOD
DEMOD
122T
MOD
DEMOD
124A
124T
152A
152R
154A
DEMOD
MOD
154R
DEMOD
MOD
156
RX MIMO處理器
完全/部分-CSI
162
TX資料處理器
150
158
RX資料處理器
160
資料槽

圖 1

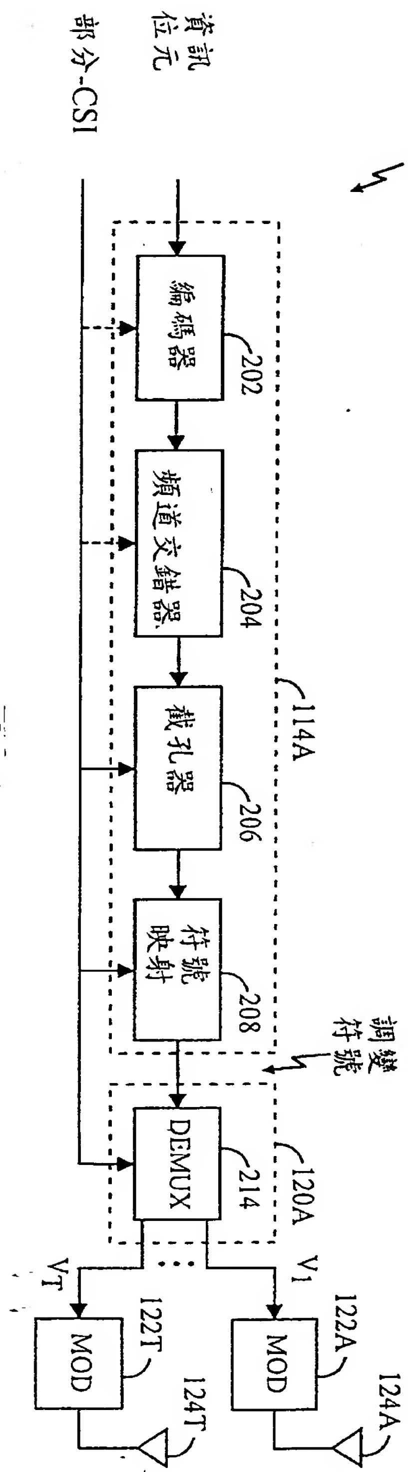

110A
資訊
位元
部分-CSI
編碼器
202
頻道交錯器
204
截孔器
206
符號映射
208
114A
調變
符號
DEMUX
214
120A
V_1
V_T
MOD
122A
MOD
122T
124A
124T

圖 2A

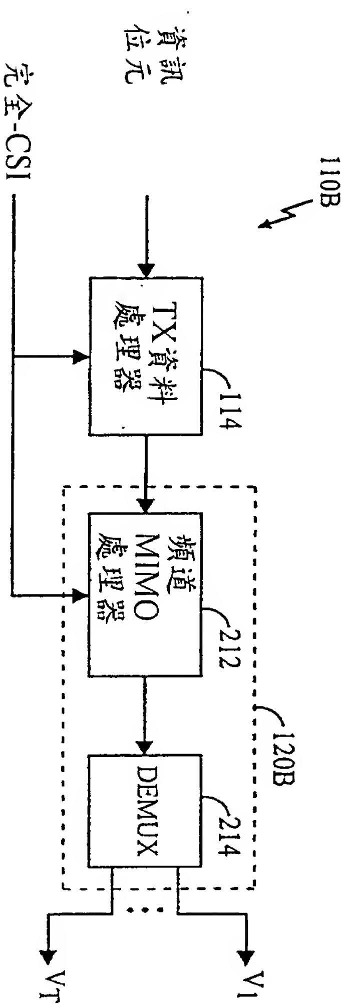

110B
資訊
位元
完全-CSI
TX資料
處理器
114
頻道
MIMO
處理器
212
DEMUX
214
120B
V_1
V_T

圖 2B

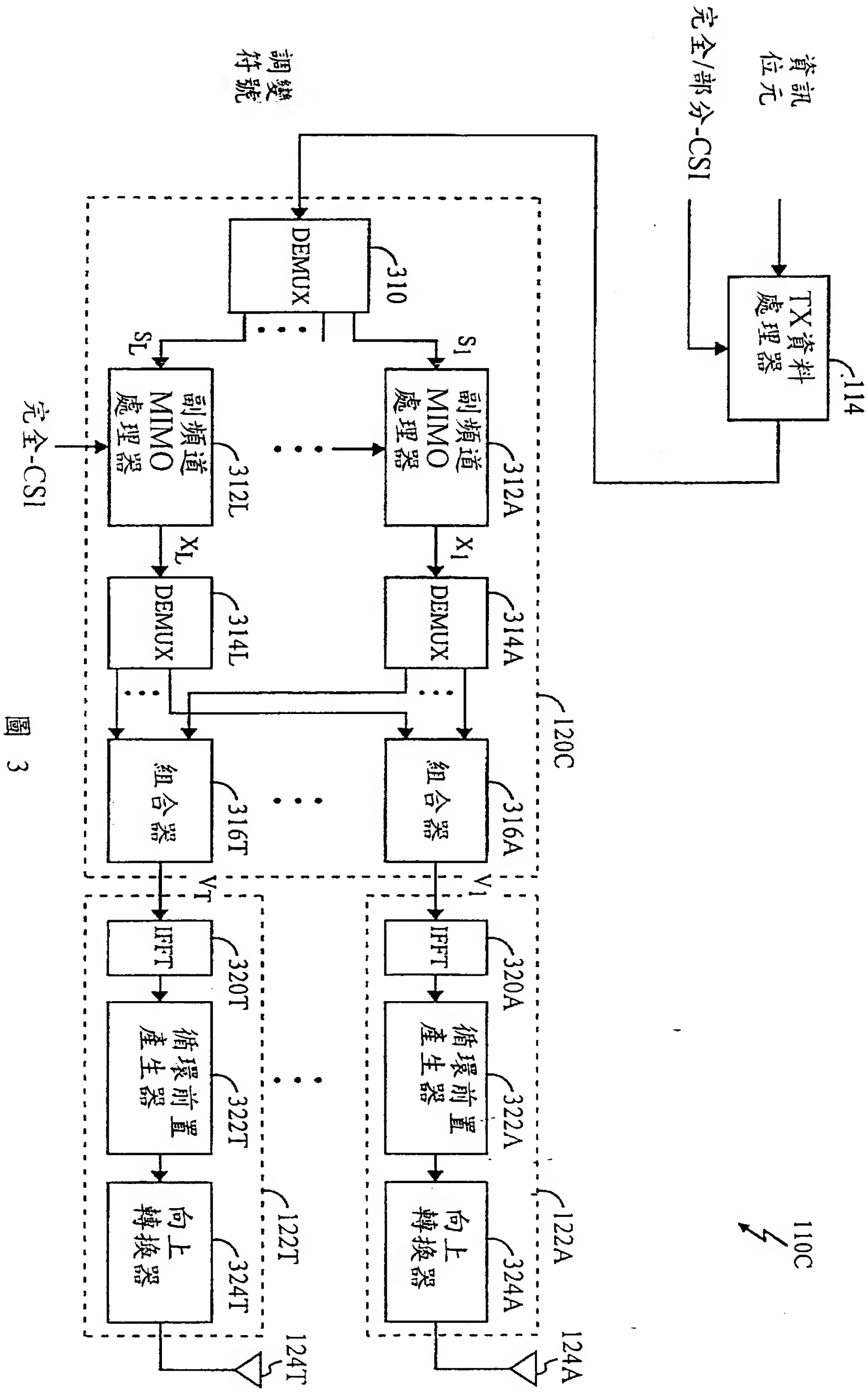

資訊位元
完全/部分-CSI
TX資料處理器
114
調變符號
DEMUX
310
S1
SL
副頻道MIMO處理器
312A
312L
完全-CSI
X1
XL
DEMUX
314A
314L
120C
組合器
316A
316T
V1
VT
IFFT
320A
320T
循環前置產生器
322A
322T
向上轉換器
324A
324T
122A
122T
124A
124T
110C

圖 3

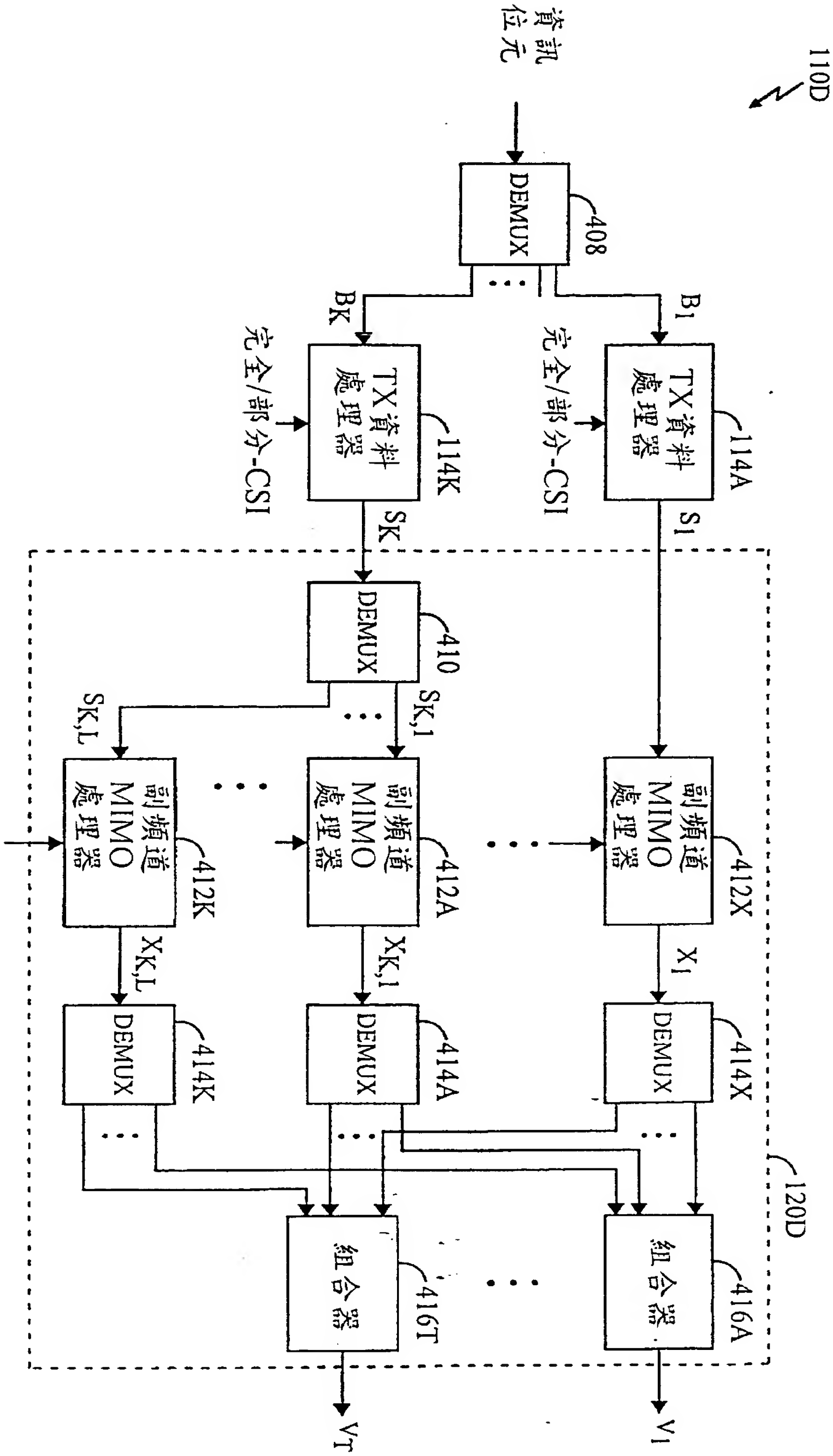

110D
資訊位元
DEMUX
408
B1
BK
TX資料處理器
114A
TX資料處理器
114K
完全/部分-CSI
完全/部分-CSI
S1
SK
DEMUX
410
SK,1
SK,L
副頻道MIMO處理器
412X
副頻道MIMO處理器
412A
副頻道MIMO處理器
412K
完全-CSI
X1
XK,1
XK,L
DEMUX
414X
DEMUX
414A
DEMUX
414K
120D
組合器
416A
組合器
416T
V1
VT

圖 4

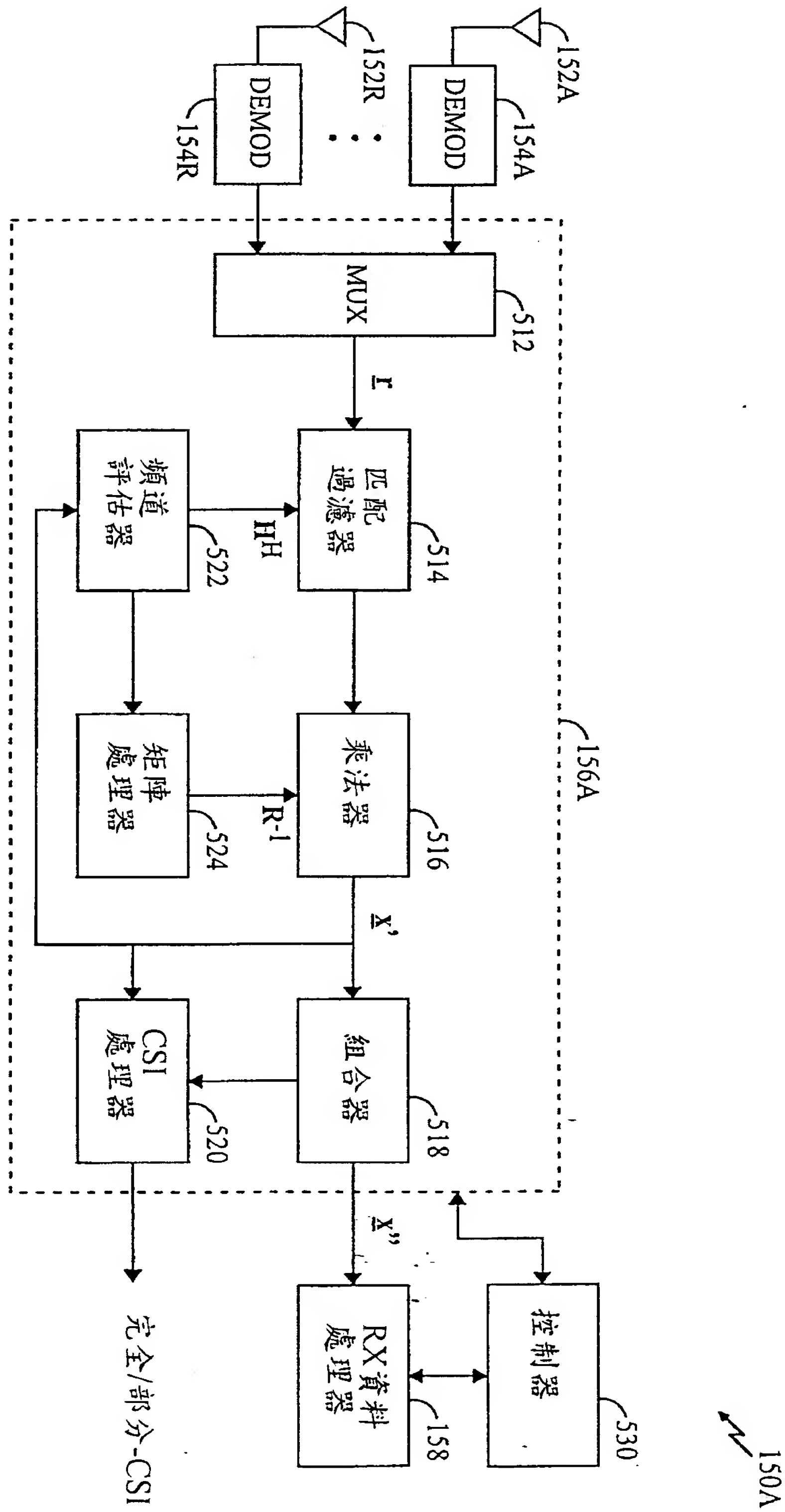

152A
152R
DEMOD
154A
DEMOD
154R
156A
MUX
512
r
匹配過濾器
514
頻道評估器
522
HH
乘法器
516
矩陣處理器
524
R-1
x'
組合器
518
CSI處理器
520
x''
RX資料處理器
158
控制器
530
完全/部分-CSI
150A

圖 5

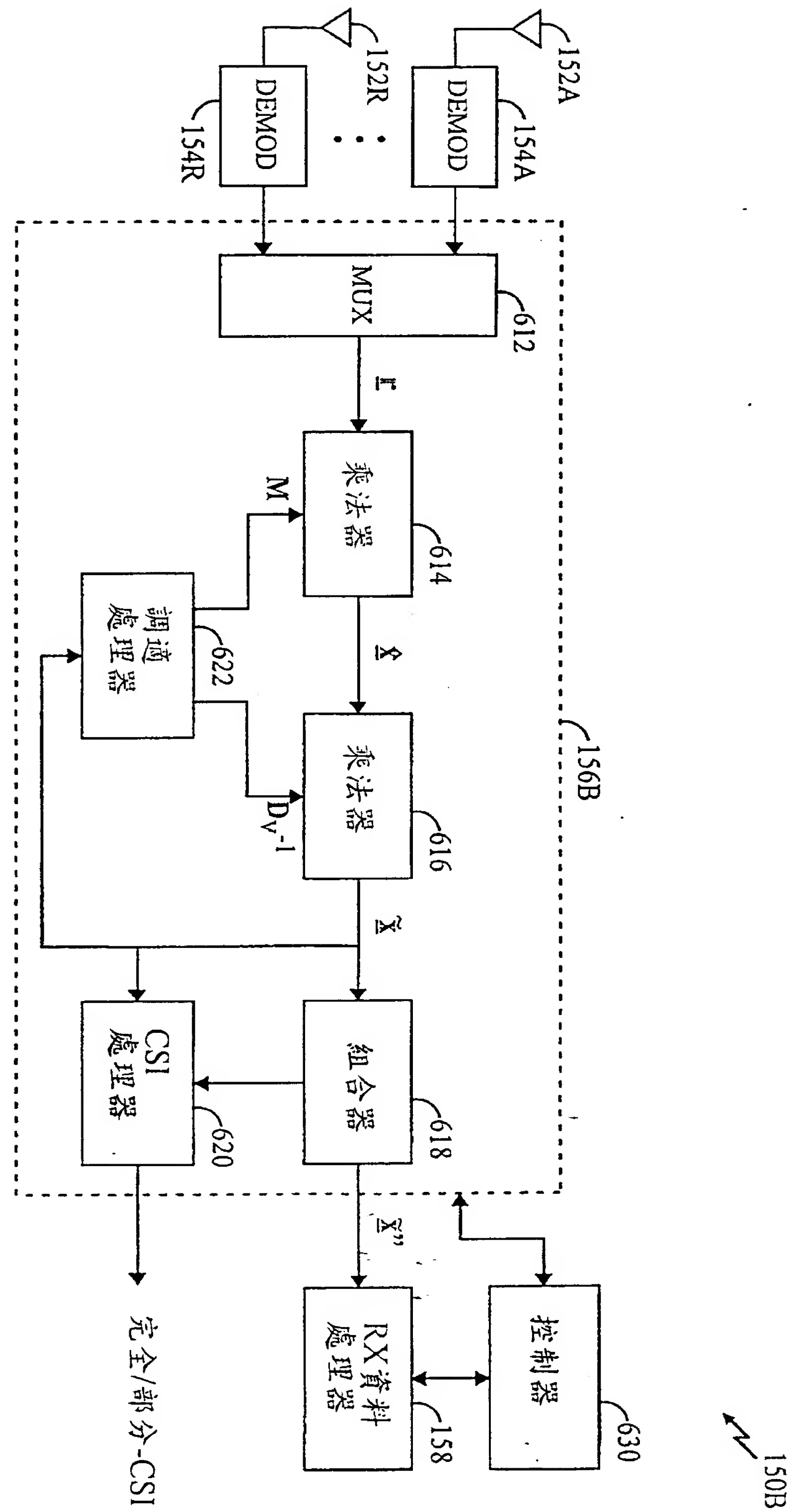

152A
152R
DEMOD
154A
DEMOD
154R
MUX
612
r
乘法器
614
M
調適處理器
622
乘法器
616
D_V^{-1}
組合器
618
CSI處理器
620
156B
RX資料處理器
158
控制器
630
150B
完全/部分-CSI

圖 6

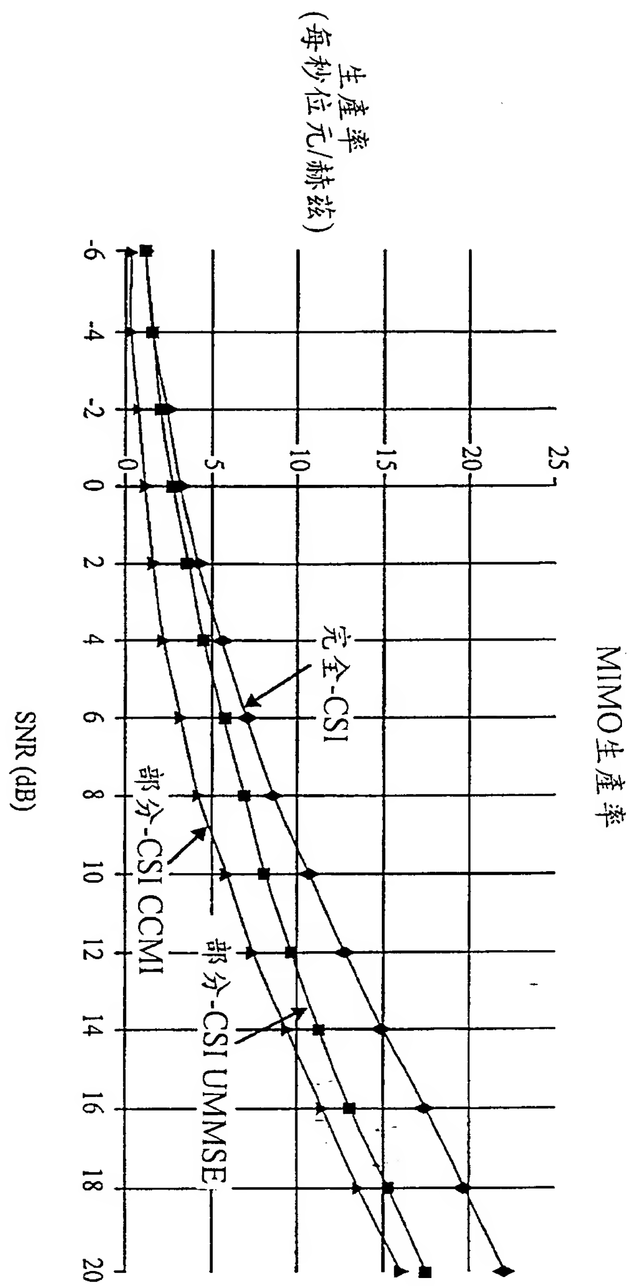

MIMO生產率
生產率
(每秒位元/赫茲)
0
5
10
15
20
25
-6
-4
-2
0
2
4
6
8
10
12
14
16
18
20
完全-CSI
部分-CSI UMMSE
部分-CSI CCMI
SNR (dB)

圖 7A

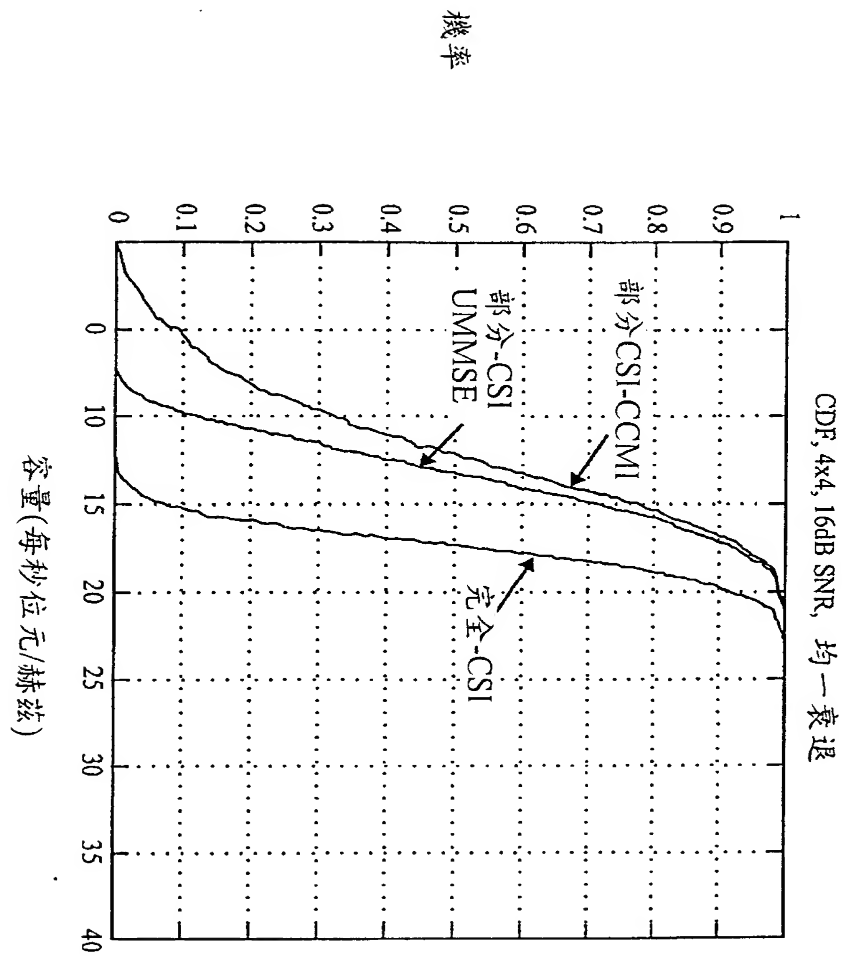

CDF, 4x4, 16dB SNR, 均一衰退
部分CSI-CCMI
部分-CSI
UMMSE
完全-CSI
機率
容量(每秒位元/赫茲)
0
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1
0
10
15
20
25
30
35
40

圖 7B

Method and apparatus for mobile platform reception and synchronization in direct digital satellite broadcast system

Patent/Publication Number 545006

Issued/Publication Date 2003/08/01

Application Date 2001/02/27

Application Number 090104466

IPC H04J-011/00;H04J-004/00;H04B-007/212;H04L-027/06;H04B-001/04

Inventor CAMPANELLA, S. JOSEPH US

Applicant WORLDSPACE MANAGEMENT CORPORATIONUS

Priority Number 20000229 US 20000185701P 20000818 US 20000640686

Abstract

A satellite system employing time diversity and a single frequency network of terrestrial re-radiation stations is provided wherein each terrestrial re-radiation station inserts a delay into a terrestrial signal. The delay allows the time of arrival of the early time diversity signal at the center of terrestrial coverage to coincide with the arrival of the corresponding late time diversity signal, thereby improving hand-off between terrestrial and satellite signals at a receiver. The delay also adjusts for distance differences between each terrestrial re-radiation station and the satellite and between each station and the center of the terrestrial coverage region. This adjustment optimizes the TDM-MCM reception by synchronizing at the center of the SFN the phase of the MCM signals re-radiated from the re-radiating stations of the SFN. The delay also compensates for the processing delay encountered when converting a satellite LOS TDM stream into a multicarrier modulated stream for transporting the satellite LOS TDM stream to user receivers and for the diversity delay between the early and late signals.

申請日期：90-2-27	案號：90104466
類別：H04J 11/00, H04J 4/00, H04B 7/212, H04L 27/26, H04B 1/06	

（以上各欄由本局填註）

發明專利說明書

545006

一、發明名稱	中文	直接數位衛星播送系統中之行動式平台接收和同步方法及裝置
	英文	Method and Apparatus for Mobile Platform Reception and Synchronization in Direct Digital Satellite Broadcast System
二、發明人	姓名（中文）	1. S. 約瑟・坎帕內拉
	姓名（英文）	1. S. Joseph Campanella
	國籍	1. 美國
	住、居所	1. 美國馬里蘭州蓋瑟士堡市維特斯頓圓環18917號
三、申請人	姓名（名稱）（中文）	1. 沃德斯帕管理公司
	姓名（名稱）（英文）	1. Worldspace Management Corporation
	國籍	1. 美國
	住、居所（事務所）	1. 美國華盛頓特區N. W. 北街2400號
	代表人姓名（中文）	1. 唐納・J・佛里克
	代表人姓名（英文）	1. Donald J. Frickel

本案已向

國(地區)申請專利	申請日期	案號	主張優先權
美國 US	2000/02/29	60/185,701	有
美國 US	2000/08/18	09/640,686	有

有關微生物已寄存於	寄存日期	寄存號碼

無

545006

四、中文發明摘要　（發明之名稱：直接數位衛星播送系統中之行動式平台接收和同步方法及裝置）

一種衛星系統，採用時間分集及提供一地面再發射站單一頻率網路，其中每一地面再發射站在地面信號中插入一延遲。此延遲允許地面涵蓋區域中心之先到分集信號到達時間與對應晚到分集信號之到達重合，以改善地面與接收器衛星信號之間的換手。此延遲亦可調整每一地面再發射站與衛星之間及每一站與地面涵蓋區域之間之距離差異。此調整可使TDM-MCM接收最佳化，即在SFN中心同步自SFN再發射站再發射之MCM信號的相位。此延遲亦可補償，當轉換一衛星LOS TDM流成為一多載波調變流，以傳送衛星LS TDM流至使用者接收器時，所遭遇的處理延遲及先到與晚到信號之間的分集延遲。

英文發明摘要　（發明之名稱：Method and Apparatus for Mobile Platform Reception and Synchronization in Direct Digital Satellite Broadcast System）

A satellite system employing time diversity and a single frequency network of terrestrial re-radiation stations is provided wherein each terrestrial re-radiation station inserts a delay into a terrestrial signal. The delay allows the time of arrival of the early time diversity signal at the center of terrestrial coverage to coincide with the arrival of the corresponding late time diversity signal, thereby improving hand-off between terrestrial and satellite signals at a

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四、中文發明摘要　（發明之名稱：直接數位衛星播送系統中之行動式平台接收和同步方法及裝置）

英文發明摘要　（發明之名稱：Method and Apparatus for Mobile Platform Reception and Synchronization in Direct Digital Satellite Broadcast System）

receiver. The delay also adjusts for distance differences between each terrestrial re-radiation station and the satellite and between each station and the center of the terrestrial coverage region. This adjustment optimizes the TDM-MCM reception by synchronizing at the center of the SFN the phase of the MCM signals re-radiated from the re-radiating stations of the SFN. The delay also compensates for the processing delay encountered when converting a satellite LOS TDM stream into a

四、中文發明摘要　（發明之名稱：直接數位衛星播送系統中之行動式平台接收和同步方法及裝置）

英文發明摘要　（發明之名稱：Method and Apparatus for Mobile Platform Reception and Synchronization in Direct Digital Satellite Broadcast System）

multicarrier modulated stream for transporting the satellite LOS TDM stream to user receivers and for the diversity delay between the early and late signals.

五、發明說明 (1)

詳細說明

本申請案主張於2000年2月29日提出之美國暫時申請序號60/185,701之優先權。

【相關申請案之交互參考】

相關主題揭示於美國專利申請序號09/058,663，1998年4月10日(放棄)，其主張1998年3月27日提出之美國暫時專利申請序號60/079,591之優先權；及1998年7月10日提出之國際PCT申請號碼PCT/US98/14280，每一申請案的整個內容併此說明供參考。

【發明之範圍】

本發明係關於一種方法及裝置，用以接收及同步，使用視線(LOS)衛星單獨接收之直接數位衛星廣播放系統，或LOS接收與地面再發射。

【發明背景】

目前存在之系統接收器提供地面及/或衛星數位音訊無線電服務(DARS)，主要受到阻擋，陰影及多路徑效應的影響，造成信號品質嚴重劣化，如多路徑造成之信號衰減及符號間干擾(ISI)。這些對廣播頻道與接收器的影響與位置及頻率關係密切，特別是都會區或常見衛星視線(LOS)信號阻擋之高地地形區域。

信號阻擋在攜帶式及移動式接收器來說，經常發生於發射器與接收器之間的實體妨礙。行動式接收器，例如，在通過隧道或建築物或樹林附近，即會阻礙視線(LOS)衛星信號接收。其他在多路徑信號反射與所要信號大小相近時

五、發明說明 (2)

亦會造成抵銷而無法提供服務。

在衛星直接下方的位置(以下稱副衛星點)實質有最大的LOS正視角度，而與副衛星點距離遠之位置實質有降低之LOS正視角，因此會增加其阻擋及陰影的機率。室外接近副衛星點的位置通常其LOS接收較無阻礙。故對可能阻礙LOS信號之地面再發射的需要最小。當衛星LOS正視角小於約85°時，高建築或地形高度(即30米大小)的阻礙即變得顯著。以填充空隙之地面再發射對行動無線電及靜態與攜帶無線電，達到滿足的涵蓋區域即有需要。在建築高度或地形妨礙較低的地區(即小於10米)，阻擋即並不明顯，除非與副衛星點的距離超過1400公里，如此LOS正視角即變成小於75°。與副衛星點相距6300公里，正視角降為25°，需要地面再發射衛星信號即明顯增加。

故在一或多個廣播衛星涵蓋區域內之中高緯度位置，即需要地面再發射才能達到適當的無線電接收。要成功施行接收直接LOS衛星信號行動式無線電及其同一信號地面再發射的組合，在接收地需要接近相對同步及衛星直接LOS信號與其地面網路中繼的組合。而且，在接收地亦需要各種地面站信號再發射信號中之接近同步。

【發明概述】

上述缺點之克服及許多優點可經由以下組合達成，即組合衛星直接LOS時間分集信號，或衛星直接LOS時間及空間分集信號，以及由地面站接收衛星直接LOS信號，再以地波再發射至城市及其郊區之再發射地面分集信號。如此，

直接LOS衛星時間或時間與空間分集信號可與正確延遲衛星直接LOS信號之地面再發射信號一起接收。因此，行動式接收器可行經衛星直接LOS信號普遍之地區，或地面再發射信號普遍之城市及其周圍郊外，或在兩種類型地區之間轉換，不致中斷其接收。要達成主要完整的連續，衛星直接LOS信號與再發射地面信號的到達時間須同步於10毫秒內。

根據本發明態樣，界定地面再發射站群之涵蓋範圍中心。由於多數地面再發射站每一站的發射信號係經校正，用以補償每一個地面再發射站及涵蓋範圍中心附近之間的距離差異。

根據本發明另一態樣，衛星信號自地面站再發射係經校正，以補償各地面再發射站之衛星先到信號到達時間的差異。

根據本發明另一態樣，自地面再發射站發射之信號係經校正，以補償在地面再發射站利用衛星信號產生地面信號造成的延遲。

根據本發明另一態樣，到達地面再發射站之分時多工資料流符號係經調整，使其與多載波調變/分時多工波形之多載波調變符號一致。

根據本發明另一實施例，至少一近似的涵蓋範圍中心附近係界定於地形分離之該地面再發射站之選擇數量中。以測定每一選擇數量地面再發射站與涵蓋範圍中心附近之間之個別距離差異。然後校正地面信號以補償在使用者終端

所選擇數量地面再發射站發射之地面信號到達的不同時間，此種時間差異係因個別選擇數量地面再發射站與涵蓋範圍中心附近之間的距離差異所產生。

根據本發明另一實施例，一種使用於地面再發射站之裝置，用以接收一分時多工資料流，包含符號，每一該符號對應所選擇數量之資料流位元。一處理裝置連接於接收裝置，用以定位資料流之主框前文。處理裝置將TDM資料流符號轉換為個別之OFDM副載波，以產生一分時多工/多載波調變(TDM-MCM)波形，包含多載波調變符號，各具有一選擇數量副載波用以傳送TDM波形之時序符號。處理裝置採用TDM主框前文，或另外分布於整個TDM框之獨一碼，以便使資料流符號與個別多載波調變符號中副載波之每一對應同步。

根據本發明另一實施例，每一再發射發射器再發射一高功率TDM-MCM信號，即自高度夠的塔台天線以地波傳播於城市或自山上或沿馬路，可到達1至20公里的適當距離。

本發明各種態樣，優點及新穎特性可由以下附圖詳細說明而更明瞭。

【較佳實施例之詳細說明】

衛星通訊系統可採用時間分集，或時間及空間分集組合，移走不要的阻擋，陰影，衰減多路徑。例如，一時間分集通訊系統可發射單一直接LOS資料流之先到及晚到衛星信號(即信號之一相對另一信號延遲一選擇時間)。或者，一時間分集通訊系統可經由個別之兩直接LOS資料流發射

先到及晚到信號。先到及晚到之時間期間係由因阻擋而停止服務期間來測定。此處之實驗證據可用來導出所需延遲。而且，兩直接LOS資料流可由空間分離之兩衛星個別發射，以施行空間分集及時間分集。在此兩情況中，非延遲頻道係在地面再發射發射器及/或在接收器處延遲，使先到及晚到頻道能建設性的組合。

上述直接LOS衛星分時施行可與地面再發射站網路組合，克服城市中心及都會區因建築，橋樑及隧道阻擋而經常無衛星直接LOS接收之問題。一地面網路可包含一至任何數量之站台，以達成所要的涵蓋範圍。使用直接衛星信號地面再發射，本發明可提供衛星TDM信號至多載波調變波形的變換，這對地波傳播之中心商業區及其周圍都會區的多路徑環境具有抗拒及強韌性。本發明提供裝置用以在地面再發射網路上以重複衛星信號同步及組合衛星直接LOS信號，當行徑衛星單獨涵蓋範圍，地面加強城市涵蓋範圍時，及當轉移於此兩類區域時，達成連續，不中斷的接收。

產生地面信號須將接收自衛星之TDM資料符號流變換成多載波調變波形，即藉IFFT轉換達成，其中TDM流資料符號係同步及精確的指定給個別的TDM-MCM副載波，所有地面信號頻率網路的地面再發射站皆使用相同方式。TDM-MCM波形係公知對多路徑具有抗拒性，且在視線接收遭到嚴重阻擋的地區仍能產生強韌接收。

<u>1. 經由衛星視線之行動式接變</u>

信號利用電磁波直接傳送於衛星發射器與行動式接收器之間係由以下定址。如前述，在接收器處的信號阻擋可能因發射器與接收器之間的實體阻礙而造成。另外，停止服務可能因信號衰減，抵銷及載波相位擾動。行動式接收器，例如，在通過隧道或行經建築物附近或阻礙視線(LOS)信號接收之樹林時，遭遇實體阻礙的阻擋。另一方面，停止服務可能因信號抵銷，衰減及載波相位擾動造成，即當多路徑信號反射干擾大到與所要信號相當時即會發生停止服務。

衛星通訊系統可採用單獨時間分集，單獨空間分集，或時間與空間分集一起，移開直接視線阻擋，陰影及多路徑衰減等不要之影響。例如，如圖1a示，單獨時間分集衛星通訊系統10可藉衛星14發射單一直接LOS資料流12同一信號的先到及晚到特別式樣(即，晚到信號係先到信號延遲一選擇時間之複製。)另外，如圖1b示，單獨時間分集通訊系統10可藉衛星14，發射一LOS資料流18僅傳送先到信號或另一LOS資料流僅傳送晚到信號。

一衛星通訊系統組合空間與時間分集，如圖2示。兩直接LOS資料流16與18可由空間分離之兩衛星14與20，其距離足以施行空間分集，個別加以發射。時間分集可由各資料流之先到及後到相伴信號的混合傳送，或由一資料流傳送所有先到信號及由另一資料流傳送所有後到信號。

圖1a，圖1及圖2任何一種系統組態，非延遲信號(即先到衛星信號)係在接收器22延遲，使其能與其後到相伴信

號同步組合成一信號。以下說明執行此種組合之最大機率方法。

圖1a，圖1b及圖2之信號宜廣播載有個別廣播節目之頻道(BCs)。個別廣播節目係指定給兩廣播頻道。一廣播頻道載有未延遲之廣播節目(即所稱之先到)。第二頻道載有相同廣播節目，但有延遲(即後到)。這些先廣播頻道可視為載有相伴信號，其中一頻道係載送於一傳送載波，另一頻道則同時載送於相反的傳送載波上。

參考圖1a之衛星信號12，一系統10接收器僅採用一衛星14之直接LOS流作為時間分集，僅需接收一時間分集多工(TDM)載波來施行此工作模式。要這樣做，接收器即使用一RF部來接收TDM載波。在此種情況，每一行動式廣播節目，兩直接LOS行動式接收廣播頻道係在同一TDM流中送出。TDM載波框內每一廣播頻道符號係與其他廣播頻道一起時間分集多工33。一廣播頻道載有一先到信號，及另一廣播頻道載有一後到信號。此程序提供接收器22之時間分集，例如，以提供交通工具沿公路移動之動態阻擋事件中，一連續不中斷的接收功能。

參考圖1b之衛星信號16與18，一系統10回復兩TDM流27及多工，並解碼這些TDM流之適當廣播，一接收器22係用來接收並處理兩TDM衛星載波。要如此做，接收器22須有一RF部能接收兩衛星TDM載波。使用具有足夠頻寬之單一RF部接收兩RFTDM載波。此種設計特別可應用在兩TDM載波配置於其頻譜內時，使其互相相鄰。但兩載波也可能出現

須在其頻譜位置中分離以產生單一RF部的情況。在此種情況中，兩分離及獨立RF部須正確定位並施行，才能接收到兩載波。此種一RF部的配置稱為單臂衛星接收器，而兩RF部的配置則稱為雙臂衛星接收器。

先到及後到信號之間的時間及期間係由要避免停止服務的期間測定。停止服務期間係由阻擋物的分布與大小來測定。在城市中，阻擋物大部分係各種高度的建築物及街道的阻礙。在郊外地區，阻擋物可能為樹林的側擋及覆蓋公路或鄉間小路。在兩種情形中，橋與隧道必須要考慮。以下將參考圖3說明以調查資料適當選擇城市及公路的延遲數值。

先到及後到信號之間的延遲時間宜為系統參數，此參數係LOS阻擋物之實體分布及車輛速度的函數。車輛以一般速度(30至60mph)行經典型郊外公路之延遲數值選擇係選擇夠涵蓋阻擋物分布之長度。延遲數值宜有足夠期間來消除97至99%所遭遇之阻擋，但不可過長，甚至涵蓋接收器構造(以避免使接收器複雜化及/或成本超出市價)。此種阻擋期間的例子，一車輛以30mph通過50呎寬橋下為例，LOS至衛星的阻擋為1.136秒，其後到信號的延遲至少等於此數值。

郊區公路的阻擋量測，如Lutz et al。，"Land Mobile Satellite communications-Channel Model，Modulation and Error Control"，於1986年5月12-16日於ICDSC-7國際會議期間數位衛星通訊所做的報告。利用其中資料繪出

阻擋對衰減深度餘裕的比例圖，例如橋，路邊構造，建築物及樹林的混合阻擋。資料如圖3所示，一衰減深度餘裕12dB，延遲時間在2至8秒之間。衰減深度餘裕係指自衛星到達信號準位與不可接受信號之間的差異。故若衛星信號夠強達到衰減餘裕12dB，由圖3知，6至8秒的延遲時間可提供時間分集接收的近似最大利益。

另外加強行動式情況衛星信號接收的裝置係一交織器。此交織器的目的係在排除位元叢波或符號錯誤，即多路徑扁平衰減及/或長阻擋造成之傳輸畸變，以消除前向錯誤校正編碼器30及其互補最大機率編碼器28組合之錯誤校正作用。即透過重新排序發射器訊息位元或符號的發生時間，以便將其隨機及均勻分布於等於交織器期間之時窗中。如此可使輸入訊息相鄰位元或符號彼此儘可能分離。若交織訊息位元或符號在其傳送至接收器時發生錯誤叢波，在接收器處之互補反交織器可回復散布於整個交織器時窗之錯誤位元或符號之原順序，使其以錯誤位元短叢波隨機分布出現於FEC解碼器，由FEC解碼器容易的加以校正。在FEC編碼器及解碼器的組合中使用此種交織器，可組成由上述系統送出之訊息或信號，端對端傳送，所使用的處理組件。此交織器係保持定位於發射器24FEC編碼器30之後及接收器22FEC解碼器28之前。其時窗期間範圍在一至多數TDM框之間。

交織器可以使用交錯交織器式樣。一交錯交織器包含一對交織器，工作於一對訊息位元流上，使每一交織器載有

每一訊息流一半的位元。訊息流位元係假性隨機及均勻分開及排序。例如，交織器輸入接受成對之訊息流。交錯交織器作用於位元以產生兩輸出交錯交織器流。交織器可使每一輸入訊息流之位元，以假性隨機方式，在兩輸出交錯交織流之間分開。而且，位元在每一對交錯交織器流中，彼此要儘可能分開遠一點。每一交錯交織器流傳送每一輸入訊息內含的一半。每一流係傳送於不同及變化的路徑中。當與一母迴旋編碼器組合使用時，其輸出即穿入兩訊息流中形成交錯交織器的輸入，及一Viterbi編碼器(利用迴旋編碼器匹配母編碼器)，輸入至母迴旋編器之訊息位元流即以最大機率方式回復於Viterbi解碼器的輸出端。此程序可消除叢波傳送位元錯誤，即行動式接收器在直接LOS衛星接收路徑所遭遇到之阻擋，陰影及多路徑衰減所造成者。

為最佳化行動式接收，晚到廣播信號及延遲先到廣播信號係儘量精確調整，使其與對應符號重合。延遲接收之先到廣播信號之數值與晚到廣播信號在發射器24之延遲34相同有助於其調整。圖1a與1b各說明有關之端對端原理。在接收器22處，兩廣播信號的符號與符號對齊係利用一固定延遲26使其儘可能精確，即使先到信號之對齊在少於一半廣播信號框周期的一半以內，然後以可變延遲調整，將服務控制標題(SCH)前文同步於先到及後到廣播信號的符號。SCHs係說明於共同授權美國專利申請序號09/112,349中，於1998年7月8日提出申請，併此供參考。此種先到及後

到廣播信號的符號對齊可讓到達接收器Viterbi解碼器28之後信號的符號達成最大機率組合。

先到及後到信號的最大機率組合係在發射器24處自迴旋編碼器30導出並將其輸出分成先到及後到時間分集信號。一公知穿入程序可完成分開，如圖示32所示。較佳之穿入含選擇先到信號迴旋母編碼位元的一半，及後到信號位元的另一半。組成每一半之精確位元係以最佳化整個端對端位元錯誤性能的方式加以選擇。在接收器處，利用軟決定Viterbi解碼器適當同步廣播流先到及後到部分的軟決定重組合可達成最佳化的最大機率組合。此再組合利用信號雜訊比預測每一再組合位元，以產生最大機率組合結果。

或者，利用相對簡單的切換取代先到及後到廣播信號的最大機率組合。在此情況，接收器22切換於先到及後到廣播信號之間。接收器22宜輸出後到廣播信號，除非後到廣播信號遭阻擋。當其遭阻擋時，接收器22則切換至延遲先到廣播信號。對齊係使用適當延遲以確保接收器22切換於先到及後到廣播信號之間時不會發生時間不連續。信號僅在先到及後到廣播信號同時受到阻擋時才會失掉。此種現象僅在同時阻擋的期間超過先到及後到信號之間的延遲時間才發生。但Viterbi最大機率組合有一明顯的信號雜訊比的優勢，即與簡單切換相比約有4.5dB。

1.1施行一衛星兩直接LOS TDM流之單獨時間分集

供行動式接收之兩TDM流係自相同衛星14傳送。一TDM流載有先到廣播信號符號，及另一TDM流載有後到廣播信號

。廣播信號宜包含複數個廣播頻道(BCs)。供行動式分集接收之BCs數量可自一至全部可使用者中變動。未使用於行動式分集接收之BCs可用來供非行動式固定及攜帶式無線電的傳統非分集LOS服務使用。先到及後到BCs提供行動式接收器的時間分集，以加強移動車輛在遭遇動態阻擋環境中的連續接收使用。載送在兩TDM流中之先到及後到BCs之間的延遲時間34係系統參數，以上述載送於相同TDM流之先到及後到BCs的相同方法測定。

在接收器22處，一對BCs，一自後到TDM流及另一自延遲先到TDM流，係以上述圖1a之先到及後到廣播信號的相同方式處理。接收器22接收兩TDM載波以施行此工作模式。

<u>1.2施行利用兩直接LOS TDM流之時間及空間分集，兩空間分離衛星每一個各一</u>

供直接衛星LOS行動式接收之兩TDM流之傳送係，一流16傳送後到信號及另一流18傳送先到信號。兩空間分離衛星14及20個別之流16及18，如圖2示，用此施行空間分集與時間分集接收兩衛星14及20係在空間充分分開，以提供TDM流兩不同到達路徑。故所提供之空間分集接收，其中之一若遭阻擋，另一個亦遭阻擋的機率不大。一TDM流16載有後到BCs，另一TDM流18則載有先到BCs，以提供接收器22之時間分集，並強化移動車輛在動態阻擋環境中連續接收使用。先到及後到TDM流之間的延遲時間34係系統參數，係由上述一TDM信號解多工之先到及後到廣播信號來測定。

1.3施行使用兩直接LOS廣播頻道之時間及空間分集，兩空間分離衛星每一個各一

供直接衛星LOS行動式接收之兩廣播頻道(即一BC傳送後到信號資訊及另一BC傳送先到信號資訊)係由兩空間分離衛星14及20每一個各一送出，如圖2說明。TDM流16及18並不須專屬所有先到或所有後到信號，而是每一個皆能傳送兩個信號的組合。此可施行空間分集及時間分集接收。兩衛星14及20在空間上分離夠遠，以提供TDM流兩不同到達路徑。故可提供空間分集接收的機會，即若一路徑阻擋，另一路徑亦阻擋的機率不大。先到及後到BCs提供接收器22之時間分集，並加強行動車輛在遭遇動態阻擋環境之連續接收。先到及後到TDM流之間之延遲時間34係系統參數，係由上述一TDM信號解多工之先到及後到廣播信號來測定。

在接收器22處，成對廣播信號(即一個載送一後到信號及另一個載送一先到信號)之處理係與上述圖1a與1b所述先到及後到廣播信號之方式相同。接收器22接收兩TDM載波以施行此工作模式。空間分集係以前述時間分集相同處理電路來實施，即由最大機率Viterbi組合處理28同時施行時間分集及空間分集。或者，可使用簡易切換選擇最佳接收品質信號。

由上述，空間分集接收產生係因衛星14之先到廣播信號，衛星20之後到廣播信號，(或對調)及衛星14與20係在不同的空間位置，如圖2示。不同空間位置之達成係，例如

，利用地面同步軌道不同位置的衛星，或兩衛星在不同橢圓軌道相對赤道傾斜，及對其之恒星日相適當地定時，以提供對目標區連續的空間及時間分集涵蓋範圍。在後者的情況中，可利用高度傾斜橢圓軌道中之三或四個衛星，例如，一次使用兩個以達成高緯度之空間分集。

2. 衛星視線阻擋之接收器地面再發射

以上任何一個直接LOS衛星分集的施行可與地面再發射發射器(圖4)網路36組合，用以在無衛星直接LOS接收之市區及郊區位置，克服因建築物，橋樑，隧道造成之阻擋，保持行動式接收器廣播節目信號的無中斷接收。一地面網路36可包含一至任何數量的站台38，例如，以符合城市或公路涵蓋範圍之需要。

行動式接收亦可另有選擇，即使用衛星直接廣播傳送系統，無需時間或空間分集，僅與地面再發射網路聯結即可。此種選擇對衛星波束涵蓋範圍，例如，LOS對衛星之正視角係85°或以上且障礙阻擋分散之地區有效。在此種環境下，僅有少數較小且隔離的阻擋區才需要地面再發射。以下說明衛星與地面接收之間的切換臨限。

使用直接衛星信號地面再發射以地面網路中繼最有利，需同步及組合行動式接收器之衛星直接LOS信號。根據本發明，以下說明施行行動式分集接收之同步，使用一或多個直接衛星LOS流，有或者沒有經由地面中繼站網路36之地面再發射。在以下說明中，信號係假設利用時間分集多工加以傳送。但這並不排除使用其他如分頻多工或分碼多

工之任何多工方法組合的傳送設計。

設置有空間及時間分集之直接LOS衛星載波可使用上述方法於無阻擋及部分阻擋之郊外地區進行使用性高之通訊。但在城市中常見之低，中及高建築會嚴重阻擋LOS衛星接收。故需要一種地面再發射系統，用以強化LOS衛星接收並達成在城市及鄉下高可用性之接收。

要克服LOS阻擋，須提供地面再發射站38網路36於城市的不同位置，如圖4示。每一地面再發射站38發射一可對付多路徑干擾之波形，及中繼直接LOS衛星數位TDM流或TDM流選擇之部分(即廣播頻道)。所有地面再發射站38宜在主要相同的載波頻率上發射。其波形頻寬互相重合，即通常所稱之單一頻率網路。可使用之波形如:1)分時多工多載波調變(TDM-MCM)，使多路徑強化技術，即公知之正交分頻多工(OFDM)來傳送一TDM信號;2)相容性TDM，所發射之TDM波形含有一特別周期性數位串列順序，致能多路徑相容等化器，即經由一關聯器，一多接點延遲線及另外的信號處理電路，串列等化器接頭以便相加性再組合之個別多路徑到達者，以回復發射之波形;及3)分碼多重進出(CDMA)，其中衛星TDM波形係分成如主要速率頻道(PRCs)之組成部分，及這些部分以佔有共用頻道之多重同時CDMA信號再廣播，並在接收器處藉每一PRC單獨配置之數位碼加以個別識別及鑑別。上述PRCs說明如上述共同授權美國專利申請序號09/112,349，於1998年7月8日申請，併此供參考。一TDM廣播波形之BC可分割成如PRCs。

以下選用一使用TDM-MCM波形之地面再發射實施例。TDM-MCM波形一詞係指以直接接收自衛星之TDM波形數位符號對多載波的調變或MCM符號。此實施例之重要態樣係將TDM-MCM地面再發射波形與接收自衛星的TDM流同步。須知此同步在由衛星送出之TDM波形與任何使用於地面再發射的其他波形之間的同步，須計入衛星與地面再發射站之間及地面再發射站與接收器之間之傳播延遲差異。

2.1 施行使用TDM-MCM之地面再發射

不同衛星傳送的選擇是可能的。他們是:1)自載送無時間或空間分集信號相同衛星之一直接LOS衛星TDM流;2)自載送先到及後到廣播信號相同衛星之一直接LOS衛星TDM流;3)自相同衛星之兩直接LOS衛星TDM流(即一流載送後到BCs及另一流載送先到BCs);及4)自不同衛星之兩直接LOS衛星TDM流(即一TDM流載送後到BCs及另一TDM流載送先到BCs，或每一流載送先到及後到BCs的組合，其中每一後到BC具有一其他TDM流中之先到同伴)。

在第一種情況，沒有使用時間或空間分集，載送BCs之TDM流係由地面再發射站38利用TDM-MCM波形直接中繼加以接收。在此種情況，於其LOS衛星TDM接收中引進一延遲配合地面再發射路徑遭遇到的處理及傳送延遲。在另外的情況中，TDM流載送先到BCs係由地面再發射站38利用一TDM-MCM波形加以延遲及中繼。

經選擇及載送於TDM-MCM波形之TDM位元流宜載送與自衛星進來之相同內含。或者，由TDM-MCM僅自衛星TDM流中選

擇供行動式接收用之BCs。在後者的情況，供行動式服務接收器用之區域注入廣播頻道內含可取代其餘TDM之容量。

根據本發明，有關時間分集接收之組態，在每一地面站插入一延遲，並調整使到達地面涵蓋範圍中心之先到BC的時間與其自衛星相伴之後到BC的到達時間相重合。此延遲包括各站38與衛星之間的距離差異，及各站38與面涵蓋範圍40中心42之間的距離差異，及將LOS TDM流轉換成TDM-MCM流有關的處理延遲。

要達到地面再發射信號與地面涵蓋範圍中心42後到衛星信號的到達時間接近重合，到達地面涵蓋範圍40內及邊緣的時間差異須為最小。因此，當當離開或進入地面涵蓋範圍40時，地面及衛星信號之間的"換手"，例如，在所接收之音頻訊號不可有明顯的中斷。此相同之對齊要求，當應用於每一地面中繼站時，使各地面MCM符號的時間與相位重合到達地面涵蓋範圍中心，有最佳化之行動式平台的接收品質。在行動式接收器離開地面涵蓋範圍中心時，MCM的到達時間與相位變成分散。經由設計，分散可大到成為護衛時間，插入至MCM符號典型為60微秒的周期內，允許自涵蓋範圍中心的分離距高達9公里。

根據本發明，每一再發射發射器以地波傳播自高功率的發射器再發射其TDM-MCM。發射之功率準位在稀疏阻擋的小涵蓋範圍為0dBW，在如商業市中心之大涵蓋範圍為40 dBW。信號係自高度足以克服阻擋環境的塔發射，其中要

納入山丘與高建築物自然地形的考慮。而且，信號係自塔上正確瞄準窄波束天線沿道路發射，其高度以地波足以到達2至16公里。

2.2 衛星LOS地面再發射信號之間的換手

換手係指當行動式車輛在LOS TDM接收衛星及地面TDM-MCM接收地面SFN之間的轉移。有兩種可能執行換手的方法。兩者皆已在前述說明過。一"換手"技術可經由調整對齊地面及衛星BCs的BC服務控制標頭前文來施行(即對齊其關聯波尖)。此程序可精確同步地面及衛星BC符號及施行其最大機率Viterbi解碼器28組合。此種施行可得到透明，無衝擊的換手。

上述技術另外可利用地面及衛星所得信號之間的切換，而不使用最大機率組合。行動式接收器調諧及收聽LOS衛星TDM載波及地面SFN再發射TDM-MCM載波兩者之一。兩載波類型傳送相同的BCs。在任何瞬間，接收器22宜選擇提供最佳BC品質之信號(即LOS TDM或TDM-MCM)。接收品質可透過各接收位元流中之位元錯誤率(BER)加以量測。切換係以BER差異完成，說明如下。

當TDM-MCM BER ≦ LOS TDM BER-△1BER時，由LOS TDM切換至TDM-MCM；及

當LOS TDM BER ≦ TDM-MCM BER-△2BER時，由TDM-MCM切換至LOS TDM

使用上述之△1BER及△2BER，可防止LOS TDM與TM之間切換時之串音。若△2BER>△1BER，TDM-MCM至LOS TDM的

切換較自LOS TDM至TDM-MCM的切換為難。這是有必要的，當進入城市涵蓋範圍時，接收器22在遭TDM-MCM捕捉時宜留在TDM-MCM。此種工作之例子，假設在加強區40，LOS TDM BER=10^{-1}及△1BER=△2BER=10^{-2}。LOS TDM至TDM-MCM的切換發生於0.01-0.001=0.009，及TDM-MCM至LOS TDM的切換再次發生，即TDM-MCM發生0.01+0.001=0.011。故LOS TDM至TDM-MCM較TDM-MCM回至LOS TDM容易。若△2BER=$4x10^{-2}$，TDM-MCM切換回至LOS TDM發生於TDM-MCM BER=.015，使其在一當選擇地面MCM時更難回至LOS TDM。其他等效方法，如信號雜訊比可取代BER用來量測品質。

2.3地面TDM-MCM傳送的施行

衛星LOS TDM流資料符號宜精確的對齊TDM-MCM資料符號內之OFDM副載波，以達到最佳SFN工作。在所說明的實施例中，每一TDM資料符號含2位元。根據本發明，完全相同的2位元指定給SFN40各地面再發射TDM-MCM中相同之OFDM副載波。此種對齊調整在每一地面再發射站38須相同的執行，因為網路任一地面再發射站的對齊調整的任何偏差，會導致TDM-MCM成為干擾，使接收品質劣化。

對齊-T資料符號至TDM-MCM波形的每一MCM符號係利圖5之說明。首先，載有接收自衛星先到BCs的TDM流TDM資料符號係依時間相鄰區塊順序排成陣列。每一TDM符號載有2位元。TDM資料符號每一區塊44含M行N列。M與N係設計參數，由TDM-MCM轉換多工器細部設計測定。最先之TDM符號填充陣列最先的列，下個最先的填充下個列，依此類推，

直至最後一列由框中最後TDM符號填充。每一區塊44係供應至反轉快速傅立葉轉換(IFFT)46之輸入端。IFFT之動作形成一含NOFDM載波之MCM符號48，即一列每一TDM資料符號一載波。每一OFDM載波係相對一加入相位參考載波之差動QPSK調變。故每一MCM符號含N+1載波。程序係依序重複於TDM資料符號區塊所有的M行，形成一完全的MCM符號框50。TDM區塊44M行形成M MCM時間順序符號48，各具有N載波加上一相位參考載波。此即組成TDM-MCM框50。每一TDM-MCM框50所載之TDM資料符號總數係M×N。須知值M=8與N=6，如圖5示，僅係用來說明。此值一般，例如，為M=960及N=116。

TDM-MCM單一頻率網路的最佳工作，自網路36每一地面再發射站38發射之每一TDM-MCM符號48，在每一MCM符號相同載波上載有區塊相同的TDM資料符號；否則，SFN36不同地面再發射站38到達接收器22之多重TDM-MCM符號48中，不會有相加性的再組合。TDM-to-MCM符號同步及對齊程序係在每一地面再發射站中以獨立但完全相同的方式執行。

MCM符號48形成TDM-MCM框50在圖6中有進一步說明。一TDM流以一符號速率R(即位元速率$B_R=2\times R$)每一符號傳送2位元至N_{TDM}符號組52之IFFT。符號宜以複數I與Q值儲存並形成行陣列之後再輸入至IFFT。一大小為2^n之IFFT46將N_{TDM}符號52轉變成N_{TDM}正交相移按鍵(QPSK)載波，產生各TDM-MCM符號，如圖6中54所示。前述I與Q值直接測定每一QPSK調變MCM OFDM載波。故每一TDM-MCM符號有N_{TDM}OFDM載波佔

有停留期間$Tsym=N_{TDM}/R$。因此，MCM符號速率$=R/N_{TDM}$。每一周期時域取樣數$=2^n$。因此，IFFT46輸出之時域MCM符號取樣率係2^n R/N_{TDM}。如圖示56，產生一護衛期間，即符號期間的分數η。此動作造成IFFT輸出的$(1-\eta)-1$時間壓縮。要組成一TDM-MCM框，在每一M_{MCM} MCM符號一次加入一框同步字元49，進一步乘積時間壓縮$(M_{MCM}+1)/M_{MCM}$，如圖示58。TDM-MCM波形頻寬即成為$R(R/S)((1-\eta)-1)(M_{MCM}+1)/M_{MCM}$。

在TDM-to-MCM符號調變中所使用之參數(即TDM流符號速率R，每一MCM符號TDM符號之N_{TDM}數，IFFT係數2^n數，護衛期間分數η，及TDM-MCM框長度M_{MCM})係經選擇以達到每一TDM框64一整數之TDM-MCM框50(圖9示)。此種選擇允許TDM-MCM框同步使用TDM主框前文(MFP)。IFFT一次接受2^n輸入係數。2^n數必須等於或大於N_{TDM}。因此，僅只有N_{TDM} OFDM副載波非零頻譜係數才能輸入至IFFT46。那些選擇之N_{TDM}值係IFFT頻譜窗中心之值。在IFFT頻譜窗邊緣未使用之IFFT係數則給予零值。

2.4TDM資料符號至TDM-MCM資料符號同步

如上述，TDM-MCM地面再發射站38宜工作於單一頻率網路(SFN)36。SFN36包含多重地面再發射站38，至少再發射部分先到衛星LOS TDM波形。所有地面再發射站以相同載波頻率帶寬發射。每一地面再發射站再廣播相同的TDM-MCM波形與其所有同伴。每一地面再發射站接收及延遲載送先到BCs之相同衛星LOS TDM信號，使載送於TDM-MCM載

波之TDM流之解調信號與載送後到BCs之衛星LOS TDM在到達SFN涵蓋範圍中心之瞬間同步。載送先到BCs之衛星LOS TDM符號精確及一致的指定給TDM-MCM資料符號的相同OFDM載波，以下以圖5與圖6說明。

定位SFN36站台38可站台數量最少的城市及其郊外之涵蓋範圍最佳化。根據本發明，在地面再發射站38引進時間延遲校正，使載送相同衛星TDM資料符號之MCM符號到達時間在涵蓋範圍中心或中心42接近同步。需要三種類型之時間延遲校正。兩種時間延遲校正與距離校正有關。一種是校正個別地面再發射站與衛星之間的距離差異，第二種係校正每一地面再發射站與SFN涵蓋範圍中心之間的距離。這兩種延遲校正的計算說如下。

第三種延遲校正係引進使TDM-MCM信號與SFN涵蓋範圍中心行動式接收器之衛星LOS後到信號同時相到達。這必須完成才能利用衛星先到LOSBCTDM信號產生TDM-MCM地面再發射信號。到達涵蓋範圍中心之後者必須與衛星LOS BC TDM信號之到達時間幾乎相同。要達到此目的，一等於先到及後到信號之間的延遲量必須用來延遲衛星之先到LOS BC TDM信號。這些延遲有些係因為在TDM至TDM-MCM轉換多工程序中造成之處理延遲。其餘則以一數位延遲線，在TDM至TDM-MCM轉換多工程序之前加至TDM流中。

可在SFN內加入一數量之"涵蓋範圍中心"，使整個城市及其郊外的接收最佳化。由於距離，群聚及阻擋的特殊情形，SFN36地面再發射站38的分組可聚集於城市及其郊外

內不同的涵蓋範圍中心。這些會影響上述前兩種校正。

3. 校正自衛星及至SFN涵蓋範圍中心距離之再發射站正時

如前述，正時校正係用來同步站台38再發射TDM-MCM信號到達SFN涵蓋範圍中心，供：

a）衛星TDM信號自衛星14或衛星14及20到再發射站台38之不同到達時間，及

b）因再發射站台38與SFN涵蓋範圍42之間距離差異產生之不同轉移時間。

正時差異可在每一再發射站台欠進，即將TDM流TDM資料符號，在記憶裝置中延遲一適當時間，然後再輸入至IFFT46。

3.1 衛星至再發射站TDM正時差異

考慮用以接收衛星TDM信號之地再發射站38網路36。一不同於90°之正視角，即直接在頭頂上，每一地面再發射站與衛星之間的距離不同。故每一地面再發射站位置與衛星之間的傾斜範圍不同，故TDM信號到達時間亦不同。而且，每一地面再發射站38與涵蓋範圍42中心之間距離亦不同。以下的情況可說明距離差異所造成之時間差異大小。

僅供說明，以SFN再發射網路36為例，其包含一數量之地面再發射站38，其地理位置選擇足以涵蓋一城市及其相關者會區。在較簡單、小及局限的阻擋拓樸中，小數量的地面再發射站已足夠。而在大、較複雜的阻擋拓樸中；需要大量的地面再發射站。

計算地面再發射站38與衛星14之間因傾斜範圍距離造成

之延遲差，其說明如圖7示。距離差異係在各站位置交叉之地表上，在衛星視線的垂直線之間來測量。最接近衛星之站台36以1表示，最遠的以m表示，其他中間以k表示。在指向任一站台k與站台m之間副衛星點方位的方向，沿地表LOS的垂直線之間的距離差異以d_{km}表示。故最遠站台m與1之間的距離為$d_{1m}=d_{max}$。注意圖7中，最遠站台係3號，最近者為1，之間有一2號。個別LOS傾斜範圍距離以△Tslantk及△Tslantmax表示。在所有站台對衛星之正視角以elv表示。而且，注意對副衛星點之方位係假設所有站台幾乎相同。因此，利用圖8所示幾何計算站台k及m之間傾斜視線距離，以下為自衛星不同到達時間的關係：

$$0 < \triangle Tslantk < \triangle Tslantmax$$

其中：

$$\triangle Tslantmax = (d_{1m} \div c) \times \cos(elv)$$

$$\triangle Tslantk = (d_{km} \div c) \times \cos(elv)$$

= 光速，米/秒

觀察其中正時校時成分$\triangle T_{correctk}$可應用於任一站台k，以表不在接收器之衛星TDM信號的到達時間差異，如下所示

$$\triangle T_{correctk} = \triangle T_{slantmax} - \triangle T_{slantk}$$

故自衛星最遠的站台36，其正時校正較小。例如，考慮d1m=dmax=18km及elv=30°的情況。在此情況，△Tslantmax=52μs。對於站台1，最接近衛星者，其校正為△Tcorrectk=△Tslantmax=52μs。對於最大距離△

$T_{correctm}=0$。任何其他中間的站台k，其$\triangle T_{correctk}$如上式表示。

對於接近副衛星點之SFN涵蓋範圍而言，每一再發射台38對衛星的方位角，每一台皆不同，故明顯的須以上述公式適當的加以校正，即站台與衛星之間的定常數傳播延遲等量線係事際上以副衛星點為中心之地表圓，及量測其各圓之間的時間差。在自副衛星之大距離及在SFN涵蓋範圍的局限區域內，其圓形可視為直線。

接著考據因衛星移動造成之時間差的變化。上述計算可應用於與衛星，地球中心及每一所考量地面台交叉之方位平面。對於一地面靜態軌道衛星，衛星軌道位置的變化很小。實務上經常維持衛星位置於指定衛星軌道位置中心側面的三維空間50哩內。在距離21,300至25,600哩處，由於地面靜態軌道衛星位置變化所造成之方位與高度變化，對上述時間校正計算的影響不大。其大小不超過135奈秒峰對峰值。同樣，亦有因時間差造成在36內的各站台位置差異。這些都不超過31奈秒峰對峰值。當兩項相加，其淨值並不超過166奈秒峰對峰值。

但是對非地面靜態軌道衛星，如飛行Tundra，Molnya，中間圓形軌道(ICO)及低地面軌道(LEO)，上述計算宜計入相對於再發射站台38的連續改變方位及衛星正視角。有關衛星通訊技術，其計算程序係上述方法的延伸。而且，對於非靜態軌道，其計算重複進行的速率，須保持LOS傾斜路徑正時錯誤在+\-500奈秒內。

3.2　護衛時間及SFN涵蓋範圍直徑自SFN36不同地面再發射站38發射之TDM-MCM信號包含TDM-MCM框50，其產生方式如上述圖5與6之說明。在所要涵蓋範圍40位置之接收器22，其多重信號包含自各再發射站台到達之TDM-MCM框。這些到達時間互相重疊，其方式如圖10說明。重疊的展幅視端視衛星至地面再發射站距離差異及再發射站至接收器距離差異而定。TDM-MCM框可相加性組合，如果其在接收器22的到達時間差不超過用來產生TDM-MCM護衛期間$\triangle T_G$的寬度。若此護衛期間寬度係$\triangle T_G$，則自SFN再發射站台複合之到達時間差必須不超過$\triangle T_G$，及其距離差異宜不超過$c \times \triangle T_G$，其中c表光速。故SFN36地面再發射站38之最大直徑配置幾乎，如圖11示，其中一地面再發射站發射器38a係與另一38b直徑相對，距離$D=c \times \triangle T_G$。故若所有地面再發射站局限於直徑$D=c \times \triangle T_G$內，則區域內或外之任何接收器的TDM-MCM框到達時間差$\triangle T_R$係$\leq \triangle T_G$。若，例如，$\triangle T_G=60$微秒，則直徑係18公里。

前述說明假設自SFN36各站台38發射之TDM-MCM框時間係經調整，使所有框到達涵蓋範圍40幾何中心42係基本上完美的對齊調整，即所有TDM-MCM框50的到達時間差基本上為零。為達成目的，自各地面再發射站之傳輸時間係根據本發明補償其兩種類型的距離差異。如前述，第一種校正係每一站台38與衛星14之間的距離差。第二種校正係站台38與涵蓋範圍40中心42的位置之間。

3.3TDM-MCM框正時校正的計算程序

以下說明在地面再發射涵蓋範圍中心完成TDM-MCM框50所需的程序。此程序宜獨立的在SFN36的各地面再發射站38執行。圖7說明SFN36地面再發射站38配置，使用與計算與公式有關之距離。程序步驟以下舉例說明。

檢視圖7有關之術語，各地面再發射站38係以指數"i"表示，其範圍自一個最接近衛星LOS距離的i=1，到衛星LOS距離最遠的一個I=m。在涵蓋範圍中之其餘站台依LOS距離增加，在1及m之間以升序編號。水平距離差，d_{im}，即平行通過各站台I及通過站台m之間，可以加以測定。注意其平行線係垂直於各站台之衛星LOS。而且注意圖7例中，m係對應站台3。

水平距離差d_{im} 係經由乘積一正視角的餘弦函數而轉換成LOS距離差。距離D_{ic}在各站台I與涵蓋範圍c42中心之間即可量測。各站台I未校正的正時△ti係由以下測定:

$$\triangle ti=[D_{ic}+ d_{im}*\cos(elv)]/s$$

其中elv係衛星正視角及s係光速

上述公式可計算SFN各再發射站台。最小△ti以$\triangle ti_{min}$表示，係下一個要測定的。各地面再發射站之校正正時△Ti之測定如下:

$$\triangle Ti=\triangle ti-\triangle ti_{min}$$

校正正時△Ti可應用於各地面再發射站I，對齊到達時間，使其在SFN涵蓋範圍中心所有TDM-MCM框中達成一零偏置。應用此種正時校正可使整個TDM-MCM地面再發射SFN工作最佳化。m=3情況的取樣計算可說明本發明原理，其中

d_{n3}係站台n至沿方位到衛星最遠站台的水平距離，及D_{cn}係站台n自涵蓋範圍中心的距離。

正時校正應用於各再發射站台

$\angle Elv=30°$

$d_{13}=18km$	$D_{1c}=15km$	$\triangle t1=102\ \mu s$	$\triangle T1=32\ \mu s$
$d_{23}=15km$	$D_{2c}=10km$	$\triangle t2=76.6\ \mu s$	$\triangle T2=6.6\ \mu s$
$d_{33}=0km$	$D_{3c}=21km$	$\triangle t3=70\ \mu s$	$\triangle T3=0\ \mu s$

上述校正可補償衛星及各再發射站之間的距離差，加上各再發射站及SFN涵蓋範圍中心之間的距離差。另外，亦必須在各站引進一延遲以補償自衛星之先到及後到信號之間的偏置及TDM-MCM轉換多工器的處理延遲。在各站引進的總延遲必須使自衛星之後到信號與經由各地面再發射站傳送信號之間有準確的重合。故若先到及後到信號之間的延遲係以T_{EL}表示及處理延遲以$\triangle T_P$表示，則各站i總延遲ΣTi為：

$$\Sigma Ti=T_{EL}-\triangle T_P-\triangle Ti$$

例如上述的考慮及假設$T_{EL}=5$秒，$\triangle T_P=0.2$秒，各站總延遲為

$\Sigma T1=5.0-0.2-32.0\ \times 10^{-6}$
$\Sigma T2=5.0-0.2-6.6\ \times\ 10^{-6}$
$\Sigma T3=5.0-0.2$

雖然本發明係以其較佳實施例說明，須知本發明並不限制於其細節。一般技藝人士可能有各種修改及代替。所有

這些代替應仍屬本發明申請專利範圍所界定之範疇內。

【元件編號說明】

10	衛星通訊系統
14	衛星
16，18	資料流
20	衛星
22	接收器
24	發射器
26	固定延遲
27TDM	流
28	解碼器
30	錯誤校正編碼器
33	時間分集多工
34	延遲時間
36	網路
38	地面再發射站
40	涵蓋範圍
42	涵蓋範圍中心
44	區塊
46	傅立葉轉換
48	MCM符號
49	同步字元
50	MCM符號框
52	NTDM符號組

五、發明說明 (30)

54　　TDM-MCM符號
56　　護衛期間
58　　MCM框時間壓縮
64　　TDM框

圖式簡單說明

圖1a與1b各說明根據本發明實施例構造之以衛星傳送時間分集信號之廣播系統。

圖2說明根據本發明實施例構造之以兩衛星傳送時間與空間分集信號之廣播系統；

圖3說明衰減期間對衰減深度並識別使用時間分集接收最佳化之延遲；

圖4說明根據本發明實施例構造之地面TDM-MCM單一頻率網路(SFN)；

圖5說明根據本發明實施例之TDM符號與MCM副載波的同步；

圖6說明根據本發明實施例TDM符號至MCM副載波調變；

圖7說明根據本發明實施例計算衛星與地面再發射站之LOS延遲差異及地面再發射站與SFN中心之間的延遲差異；

圖8說明水平距離至LOS距離的變換，供圖7說明之TDM-MCM框正時使用；

圖9說明根據本發明實施例將TDM框分割成MCM框；

圖10說明根據本發明實施例調整具有選擇直徑SFN多數站發射之TDM-MCM框；

圖11說明根據本發明實施例SFN地面再發射站配置之最直徑；

所有圖示中相同組件及零件使用相同元件編號。

六、申請專利範圍

1. 一種同步方法，用以同步分時多工或TDM資料流中TDM符號之選擇數量，使其等於分時多工/多載波調變副載波或TDM-MCM波形中TDM-MCM符號之數量，包含步驟：

定位主框前文(MFP)碼及該TDM流一分配同步順序之一，該TDM資料流具有至少一TDM框，包含該一MFP碼及一分配同步順序及複數個該符號，該MFP碼及分配同步順序之一可用於定位該TDM資料流內之該TDM框；

產生一陣列使用該TDM框該符號，該陣列包含一第一數之行及一第二數之列；及

產生TDM-MCM符號對應於使用一反轉快速傅立葉轉換(IFF)與該陣列之該第一數，各該TDM-MCM符號具有該載波該第二數，即對應該列個別之該TDM符號，該TDM-MCM符號之該第一數對應於一TDM-MCM符號框。

2. 如申請專利範圍第1項之方法，其中該產生步驟包含步驟填充該陣列，提供該TDM框最先到達之TDM符號至該陣列該列最先產生的一個，及繼續順序填充該列直至該列最後一個填完該TDM框該TDM符號最後一個。

3. 如申請專利範圍第1項之方法，其中該TDM資料流包含複數個TDM框，該TDM-MCM符號框具有基本上與該TDM框相同的停留期間。

4. 如申請專利範圍第1項之方法，其中該產生步驟包含步驟同步該TDM-MCM符號框該TDM-MCM符號成為該TDM資料流該符號之一的分數。

5. 如申請專利範圍第4項之方法，其中該TDM-MCM符號框

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中之該TDM-MCM符號數係一整數。

6.如申請專利範圍第1項之方法，另包含步驟提供該TDM-MCM框各該TDM-MCM符號一護衛期間，TDM-MCM符號周期對應每秒該TDM符號數分割之該第二數，該護衛期間係小於該TDM-MCM符號周期。

7.如申請專利範圍第1項之方法，另包含步驟提供該TDM-MCM波形各該TDM-MCM符號框一同步字元。

8.如申請專利範圍第1項之方法，另包含步驟:

提供該TDM-MCM框各該TDM-MCM符號一護衛期間，TDM-MCM符號周期對應每秒該TDM符號數分割之該第二數，該護衛期間係小於該TDM-MCM符號周期;

提供該TDM-MCM波形各該TDM-MCM符號框一同步字元;及

壓縮各該MCM-TDM符號以補償插入各該TDM-MCM符號框之該護衛期間及該同步字元，使具有該護衛期間之該TDM-MCM符號及對應該TDM-MCM框與對應該同步字元之時間配置佔據一TDM框周期。

9.如申請專利範圍第1項之方法，其中該IFFT採用一數量係數大於該符號該第二數。

10.如申請專利範圍第1項之方法，其中該TDM資料流包含複數個TDM框，該產生步驟另包含步驟指定該TDM資料流個別該TDM框之該符號給對該TDM-MCM框之該TDM-MCM符號該副載波。

11.一種供使用於地面再發射站之裝置，包含:

一接收裝置，用以接收一分時多工或TDM資料流，包含

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符號，各該符號對應該資料流中選擇之位元數；及

一處理裝置，連接該接收裝置及可操作以定位該TDM資料流主框前文(MPF)碼及一分配同步順序，該TDM資料流具有至少一TDM框，包含該MFP碼及一分配同步順序及複數個該位元之一，該MFP碼及一分配同步順序可用於定位該TDM資料流內該TDM框；

其中該處理裝置轉換該TDM資料流該符號成為個別載波，以產生分時多工/多載波調變或包含TDM-MCM符號，各該TDM-MCM符號具有一選擇數量之副載波，該處理裝置採用該MFP碼及一分配同步順序之一，用以同步該TDM資料流該符號與個別該TDM-MCM符號中該副載波對應者。

12. 如申請專利範圍第11項之裝置，其中該處理裝置採用一反轉快速傅立葉轉換(IFFT)轉換該TDM資料流該符號成為個之該副載波。

13. 如申請專利範圍第12項之裝置，其中一TDM-MCM框包含一選擇數量之該TDM-MCM符號，該處理裝置可操作以產生在該TDM資料流該TDM框一整數之該TDM-MCM符號。

14. 如申請專利範圍第13項之裝置，其中該處理裝置係可操作以提供各該TDM-MCM符號框對應該TDM框相同之該符號。

15. 如申請專利範圍第14項之裝置，其中該處理裝置係可操作以對應該TDM框該符號給該TDM-MCM符號框中之該TDM-MCM符號之個號載波。

16. 如申請專利範圍第15項之裝置，其中該裝置係用於

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一地面再發射站，可操作以接收該TDM資料流及轉換成其中該符號至個別之載波內，以產生包含TDM-MCM符號之TDM-MCM框，該處理裝置係可操作以指定該TDM資料流個別之該TDM框符號給對應該TDM-MCM框中之該TDM-MCM符號之該副載波。

17. 一種使用於地面再發射站之系統，包含:

一接收器，用以接收衛星分時多工或TDM資料流;

一轉換編碼器，連接於該接收器用以轉換該TDM資料流成為多載波調變(MCM)波形，以產生一分時多工/多載波調變或TDM-MCM信號，可強力面對地面路徑傳輸多路徑及畸形阻擋及干擾;及

一發射器，連接於該轉換編碼器用以發射該TDM-MCM信號。

18. 如申請專利範圍第17項之系統，其中該發射器係組態成可再發射該TDM-MCM信號於地面路徑約2公里與10公里之間的距離，其中衛星接收遭阻擋。

19. 如申請專利範圍第17項之系統，其中該發射器係組態成可再發射該TDM-MCM信號於地面路徑至少一城市及沿公路至所選擇距離，其中衛星接收係分別遭建築物及樹林阻擋。

20. 如申請專利範圍第17項之系統，其中複數個該系統係置於單一頻率網路之個別地面再發射站，該系統工作實質同時互相利用正時協調及同步，以達成在該單一頻率網路相關區域實質無縫之該TDM-MCM之接收。

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21.如申請專利範圍第20項之系統，其中該地面再發射站係依地形設置用以服務城市及其周圍郊外地區。

22.一種用以發射分時通訊系統中之廣播頻道之方法，其中一先到信號與一後到信號係經由至少一衛星發射，先到信號包含至少一部分廣播頻道及後到信號包含另一部分廣播頻道，後到信號對應先到信號但相對於先信號延遲一選擇之時間周期，通訊系統包含地面再發射站網路用以接收及處理先到信號以發射一地面再發射信號，此方法包含步驟：

測定該衛星及該網路各該地面再發射站之間距離之個別差異；及

校正該地面再發射信號以補償該先到信號在個別之該地面再發射站之到達時間差異。

23.如申請專利範圍第22項之方法，其中該網路係一單一頻率網路。

24.如申請專利範圍第22項之方法，其中另包含步驟：界定在該地面再發射站之選擇數量中至少一大約涵蓋範圍中心；

測定各該選擇數量該地面再發射站與該大約涵蓋範圍中心之間距離之個別差異；及

校正該地面再發射信號以補償自該選擇數量該地面再發射站發射之該地面再發射信號到達接收器之不同時間，即因該選擇數量個別之該地面再發射站與該大約涵蓋範圍中心之間所產生者。

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25. 如申請專利範圍第22項之方法，其中該校正步驟包含步驟補償將該衛星信號轉變為該地面再發射信號之延遲。

26. 一種用以發射分時通訊系統中之廣播節目之方法，其中一先到信號與一後到信號係經由至少一衛星發射，先到信號包含至少一部分廣播節目及後到信號對應於先到信號，但對應於先到信號延遲一選擇之周期時間，通訊系統包含一地面再發射站網路，用以接收及處理延遲信號以發射一地面再發射信號，此方法包含步驟：

界定地形分離之該地面再發射站之選擇數量中至少一大約涵蓋範圍中心；

測定各該選擇數量該地面再發射站與該大約涵蓋範圍中心之間距離之個別差異；及

校正該地面再發射信號以補償自該選擇數量該地面再發射站發射之該地面再發射信號到達接收器之不同時間，即因該選擇數量個別之該地面再發射站與該大約涵蓋範圍中心之間所產生者。

27. 如申請專利範圍第26項之方法，其中該校正步驟包含步驟用以補償將該衛星信號轉變為該地面再發射信號之延遲。

28. 一種用以提供一廣播節目至接收器之方法，包含步驟：

接收利用僅時間分集或時間及空間分集發射之衛星信號，該衛星信號在最大可能性組合時包含該廣播節目；

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接收一地面再發射信號包含該廣播節目並由一地面再發射站發射；

測定該最大可能性組合衛星信號及該地面再發射信號何者有最佳信號品質；

選擇該最大可能性組合衛星信號或該地面再發射信號具有最佳輸出信號品質；及

抑制自該選擇信號至該最大可能性組合衛星信號及該地面再發射信號之另一個之切換，除非滿足一選擇條件。

29.如申請專利範圍第28項之方法，其中該選擇信號品質對應位元錯誤率選擇臨限，以接收地面再發射信號。

30.如申請專利範圍第29項之方法，其中該位元錯誤率之該選擇臨限，在選擇該地面再發射信號及抑制該最大可能性組合衛星信號時，係較大於，在選擇該最大可能性組合衛星信號及抑制該地面再發射信號時。

31.如申請專利範圍第28項之方法，其中該最大可能性組合衛星信號及該地面再發射信號並不施行時間分集或時間與空間分集。

32.如申請專利範圍第28項之方法，其中接收該衛星信號該接收步驟另包含步驟延遲該衛星信號，以補償在該地面再發射站發生之延遲，即自該衛星信號產生該地面再發射信號。

33.如申請專利範圍第32項之方法，其中該衛星信號係一分時多工信號並利用分時多工/多載波調變轉變至該地面再發射信號。

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34. 如申請專利範圍第33項之方法，其中發生於地面再發射站之該延遲係對應於處理該衛星信號，即將該衛星信號自該分時多工信號轉變成該地面再發射站之分時多工/多載波調變波形。

35. 如申請專利範圍第28項之方法，其中該衛星信號係一分時多工信號，且係利用分時多工/多載波調變將其轉變成分時多工/多載波調變之該地面再發射信號，以產生該地面再發射信號，及一行動式接收器接收及回復該分時多工信號及該分時多工/多載波調變之波形。

36. 如申請專利範圍第28項之方法，其中該衛星信號係經由一衛星以一第一頻率發射，該地面再發射信號係被至少一地面再發射站經由一第二頻率發射，該衛星信號之該接收步驟及該地面再發射信號之該接收步驟係以一第一無線電頻率接收器部及一第二無線電頻率接收器部分別在至少一該接收器上執行。

37. 如申請專利範圍第28項之方法，其中該衛星信號係經由第一衛星利用第一頻率發射，該衛星信號係經由第二衛星利用一第二頻率發射，該地面再發射信號係經由至少一地面再發射站以第三頻率發射，該衛星信號之該接收步驟係由一第一接收器部及一第二接收器部分別工作於該第一頻率及該第二頻率加以執行，及該地面再發射信號之該接收步驟係以一工作於該第三頻率之第三接收器部在至少一該接收器上執行。

38. 一種用以提供一廣播節目至接收器之方法，包含步

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驟：

接收一衛星信號包含該廣播節目，該衛星信號中之一單一廣播資料流包含一先到頻道對應於該廣播節目，及一後到頻道具有至少一部分該廣播節目在其發射之前係延遲一選擇時間周期，該先到頻道及該後到頻道各具有一同步碼，該廣播資料流已由一工作於一選擇碼速率之母迴旋編碼器加以編碼；

延遲該先到頻道該選擇之時間周期；及

在工作於該碼速率之最大可能性Viterbi解碼器組中組合該後到頻道與該先到頻道，以回復該廣播節目信號，在接收該先到頻道及該後到頻道中無任何因非關聯阻擋所造成之中斷。

39. 如申請專利範圍第38項之方法，其中該廣播資料流包含行動式接收廣播節目及靜態接收廣播節目，該先到頻道僅包含行動式接收之該廣播節目。

40. 如申請專利範圍第38項之方法，另外包含接收一地面再發射站之地面再發射信號，及一第二衛星信號包含該廣播節目及提供相對該衛星信號之空間分集，該地面再發射信號及該第二衛星信號各包含至少一部分該廣播節目及該同步碼，其中該組合步驟包含步驟：

利用該同步碼以對齊該衛星信號、該第二衛星信號及該地面再發射信號；及

利用至少該衛星信號、該第二衛星信號及該地面再發射信號中之一以組合產生一輸出。

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41. 如申請專利範圍第40項之方法，其中該先到頻道及該後到頻道係分別僅指定給該衛星信號及該第二衛星信號。

42. 如申請專利範圍第40項之方法，另包含步驟接收一第二衛星信號，包含該廣播節目及提供相對於該衛星信號之空間分集，其中該衛星信號及該第二衛星信號係自一地面同步軌道中不同軌道位置發射。

43. 如申請專利範圍第40項之方法，另包含步驟接收一第二衛星信號，包含該廣播節目及提供相對於該衛星信號之空間分集，其中該衛星信號及該第二衛星信號係自三或四個不同橢圓軌道，相對於一恒星日周期地球赤道傾斜約63°，加以發射。

44. 一種用以執行於一發射器準備供接收器Viterbi迴旋解碼器之最大可能性接收之方法，包含步驟：

利用一母迴旋編碼器以一選擇編碼速率編碼一廣播節目及於發射站產生母輸出位元；

在該發射站穿透該母編碼輸出位元產生兩較高速率迴旋編碼流，以獲得一第一組穿透編碼位元及一第二組穿透編碼位元；

指定該第一組穿透編碼位元至一未延遲之先到頻道；

指定該第二組穿透編碼位元至一後到頻道；

相對於該先到頻道延遲一選擇時間周期；及

發射該先到頻道及該後到頻道，該選擇時間周期，當該行動式接收器與該行動式接收器該發射器不全接收之間因

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實體妨礙發生服務阻擋時，允許該後到信號在行動式接收器非關聯於該先到信號。

45. 如申請專利範圍第44項之方法，其中該編碼速率係R＝1／3。

46. 如申請專利範圍第45項之方法，其中該較高速率迴旋編碼流係以速率R＝3／4產生。

47. 如申請專利範圍第46項之方法，其中該產生步驟包含步驟利用該第一廣播頻道之編碼位元每18位元之8個，及該互補組該18位元之另外8個，來組成一第二廣播頻道之編碼位元。

48. 如申請專利範圍第44項之方法，其中該先到頻道及該後到頻道係在接收器組合，重製無非關聯該服務阻擋中斷之該廣播節目。

49. 如申請專利範圍第48項之方法，其中該先到頻道及該後到頻道各包含至少一同步碼，及另包含步驟：

以該選擇之時間周期來延遲接收之該先到頻道；

在每一接收之該先到頻道及該後到頻道中關聯該同步碼；

相對於接收之該後到頻道，再精確對齊延遲之該先到頻道，對齊該廣播節目中之一符號及一位元之一之寬度分數內，即使該關聯步驟所得之關聯波尖重合；及

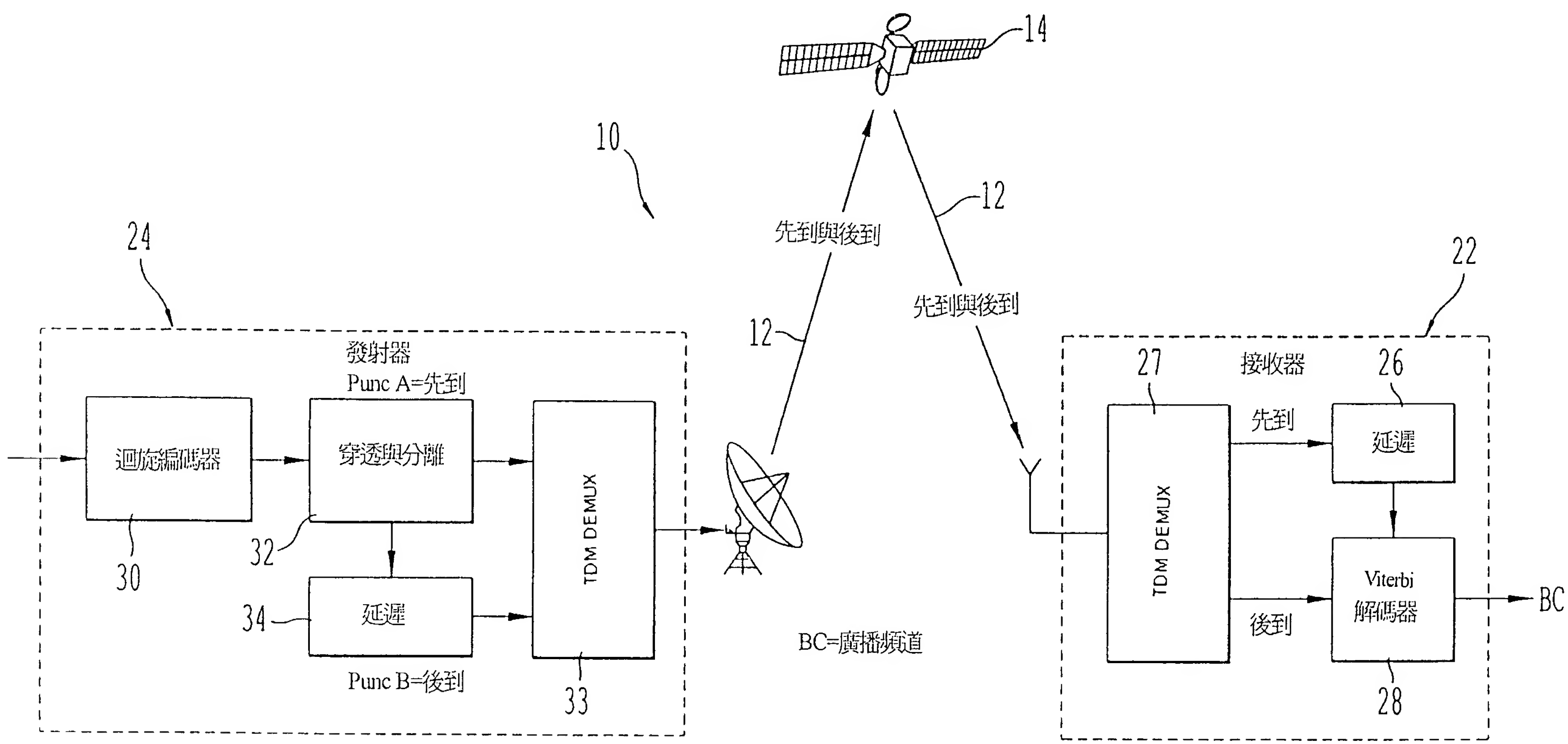
在一軟決定Viterbi解碼器中，最大可能性組合接收之該先到頻道及該後到頻道中之位元，以產生一輸出信號，無因該實體妨礙造成之無關聯服務停止。

六、申請專利範圍

50. 如申請專利範圍第49項之方法，其中該軟決定Viterbi解碼器操作於該母迴旋編碼器之該選擇編碼速率。

51. 如申請專利範圍第49項之方法，其中該選擇編碼速率係R=1/3。

545006



14
10
12
先到與後到
12
先到與後到
24
22
發射器
Punc A=先到
迴旋編碼器
穿透與分離
TDM DEMUX
30
32
34
延遲
Punc B=後到
33
BC=廣播頻道
27
接收器
26
先到
延遲
TDM DEMUX
Viterbi
解碼器
後到
BC
28

圖 1A

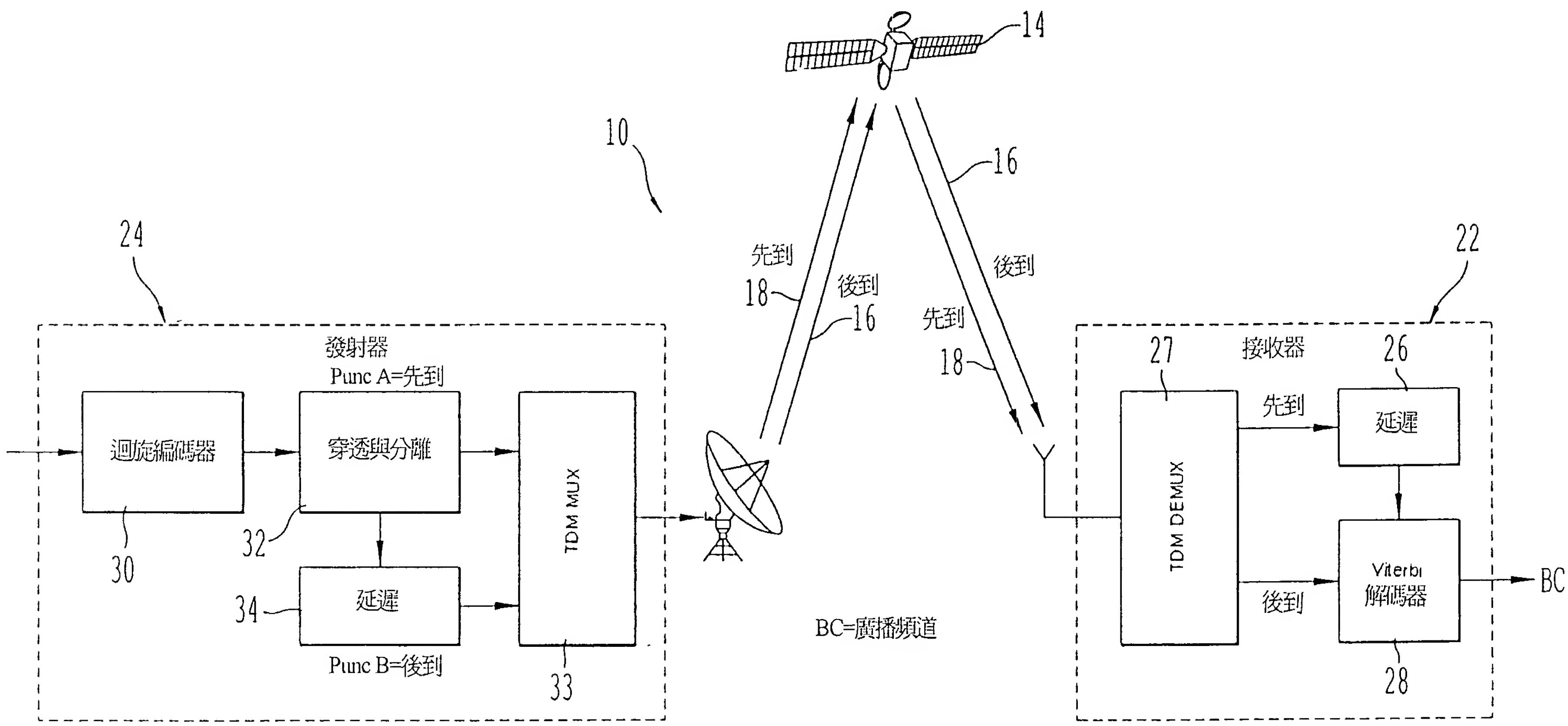
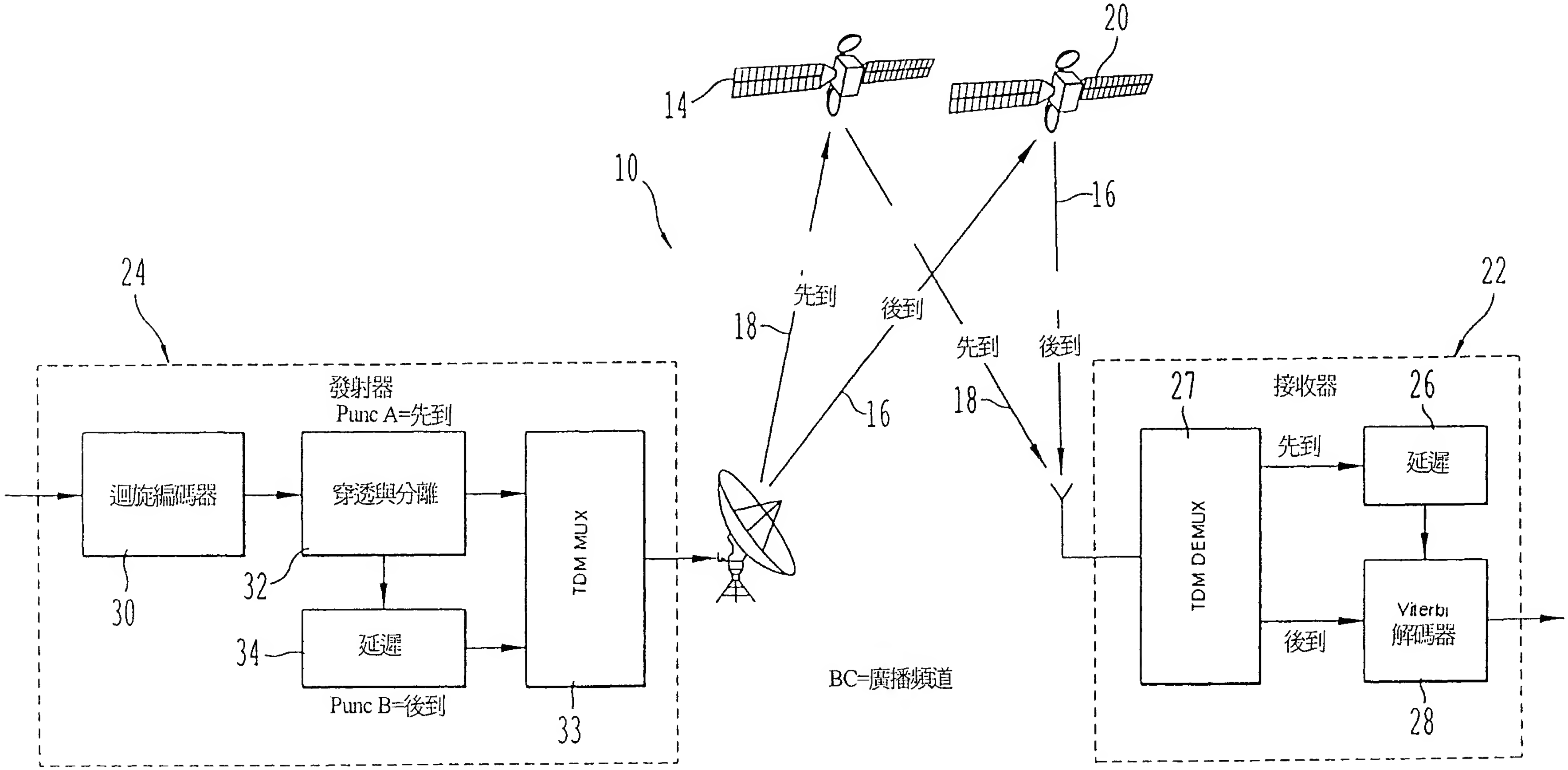

10
24
發射器
Punc A=先到
迴旋編碼器
30
穿透與分離
32
延遲
34
Punc B=後到
TDM MUX
33
先到
18
後到
16
14
16
後到
先到
18
BC=廣播頻道
22
接收器
27
TDM DEMUX
先到
延遲
26
後到
Viterbi
解碼器
28
BC

圖1B

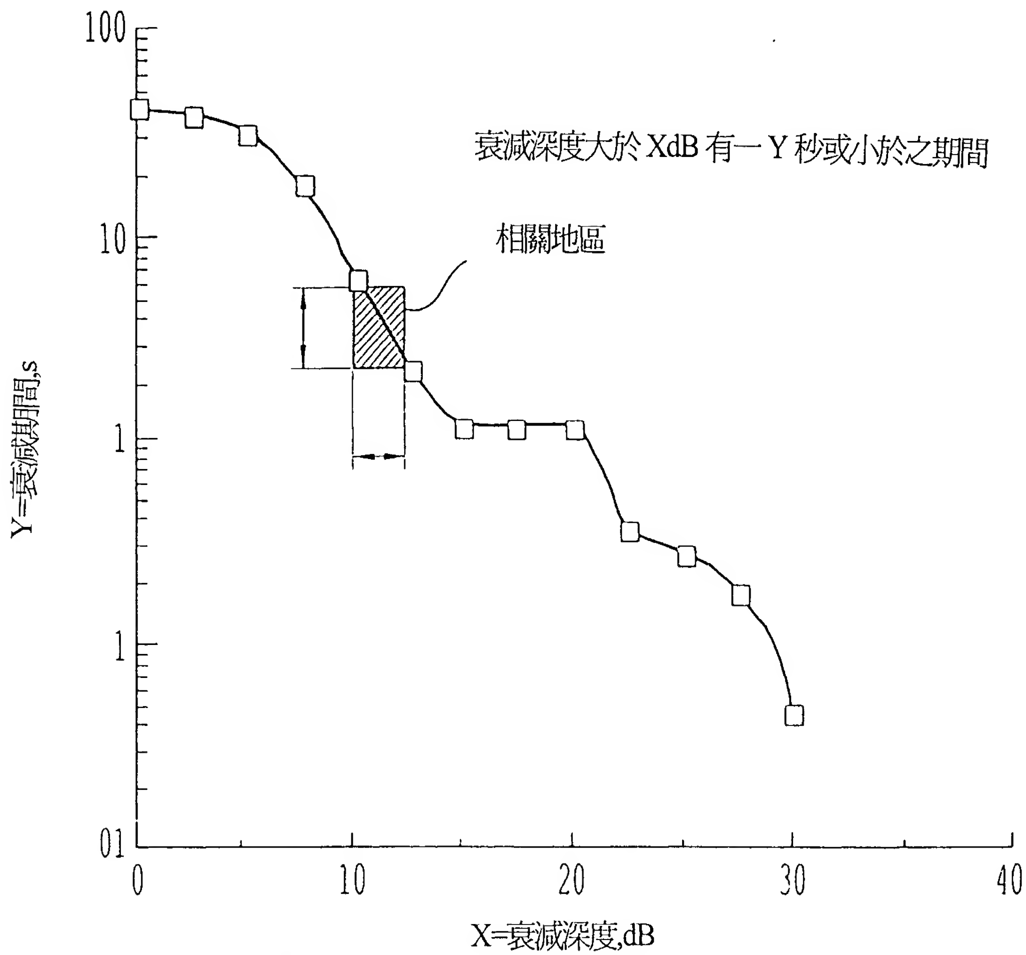


14
20
10
16
18
先到
後到
先到
後到
16
18
24
22
發射器
Punc A=先到
迴旋編碼器
穿透與分離
TDM MUX
延遲
Punc B=後到
30
32
34
33
BC=廣播頻道
接收器
27
26
TDM DEMUX
先到
延遲
後到
Viterbi
解碼器
28

圖2

545006

最大衰減期間對衰減深度圖



3

545006

92. 5. 13 年 月 日 修正 補充

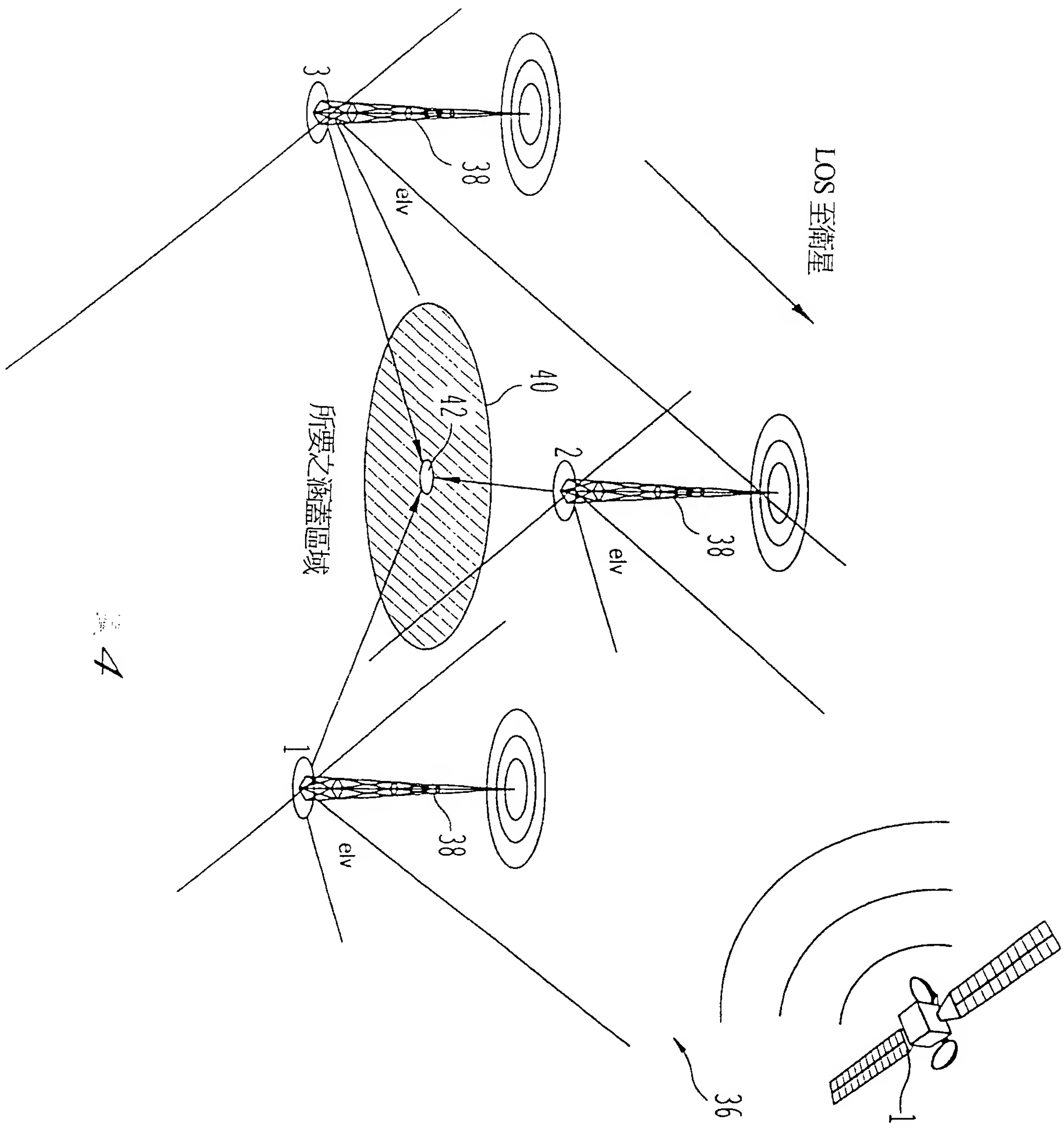
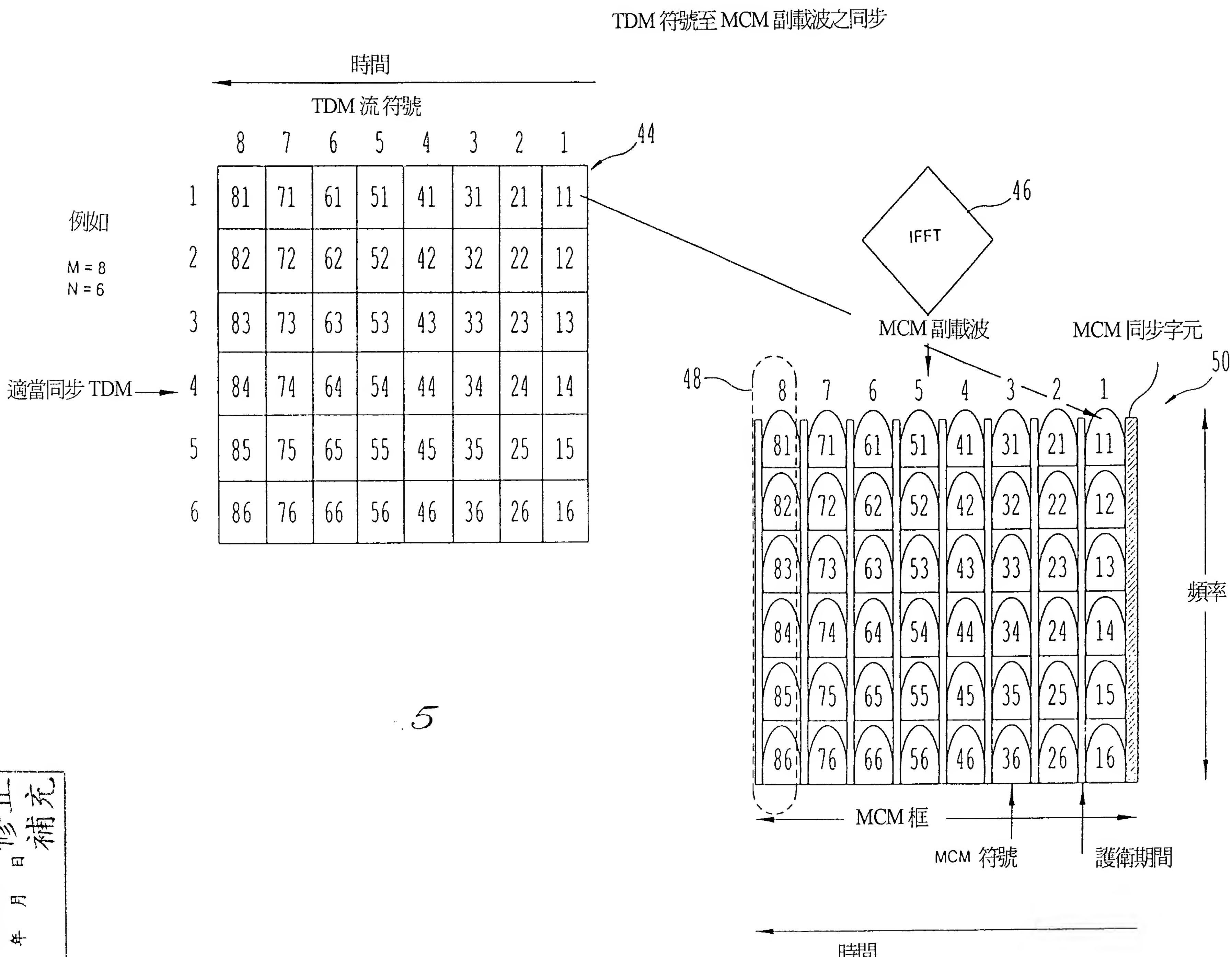
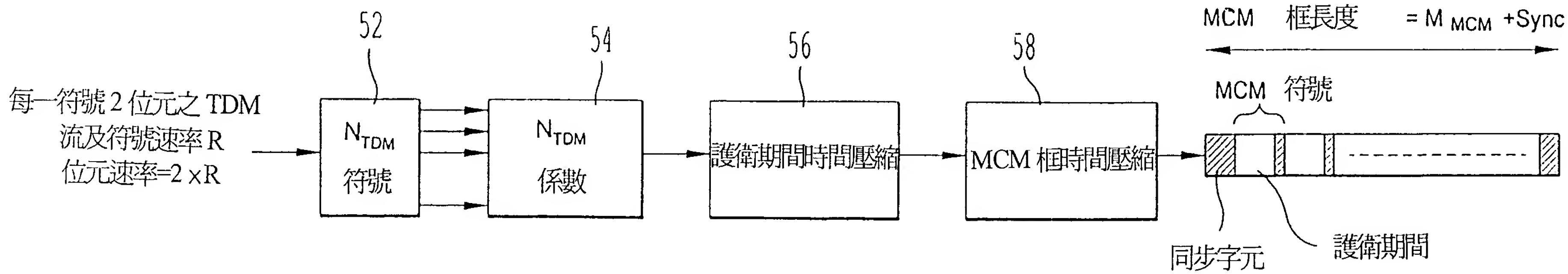


圖 4

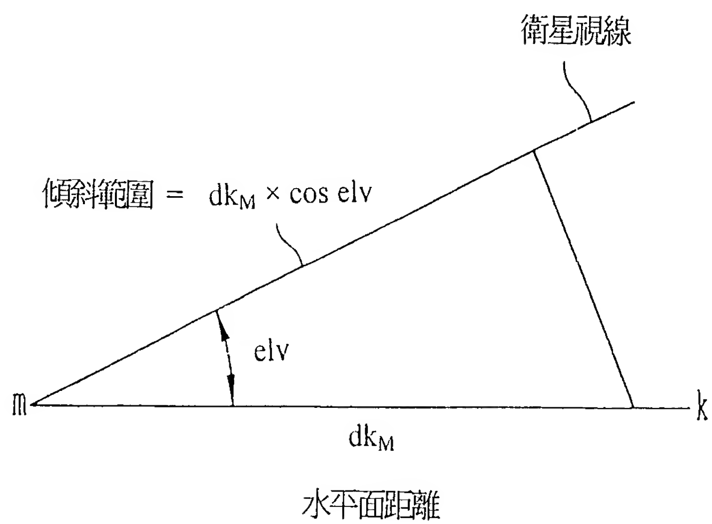

TDM 符號至 MCM 副載波之同步
時間
TDM 流符號
44
例如
M = 8
N = 6
適當同步 TDM
46
IFFT
MCM 副載波
MCM 同步字元
48
50
頻率
MCM 框
MCM 符號
護衛期間
時間

5



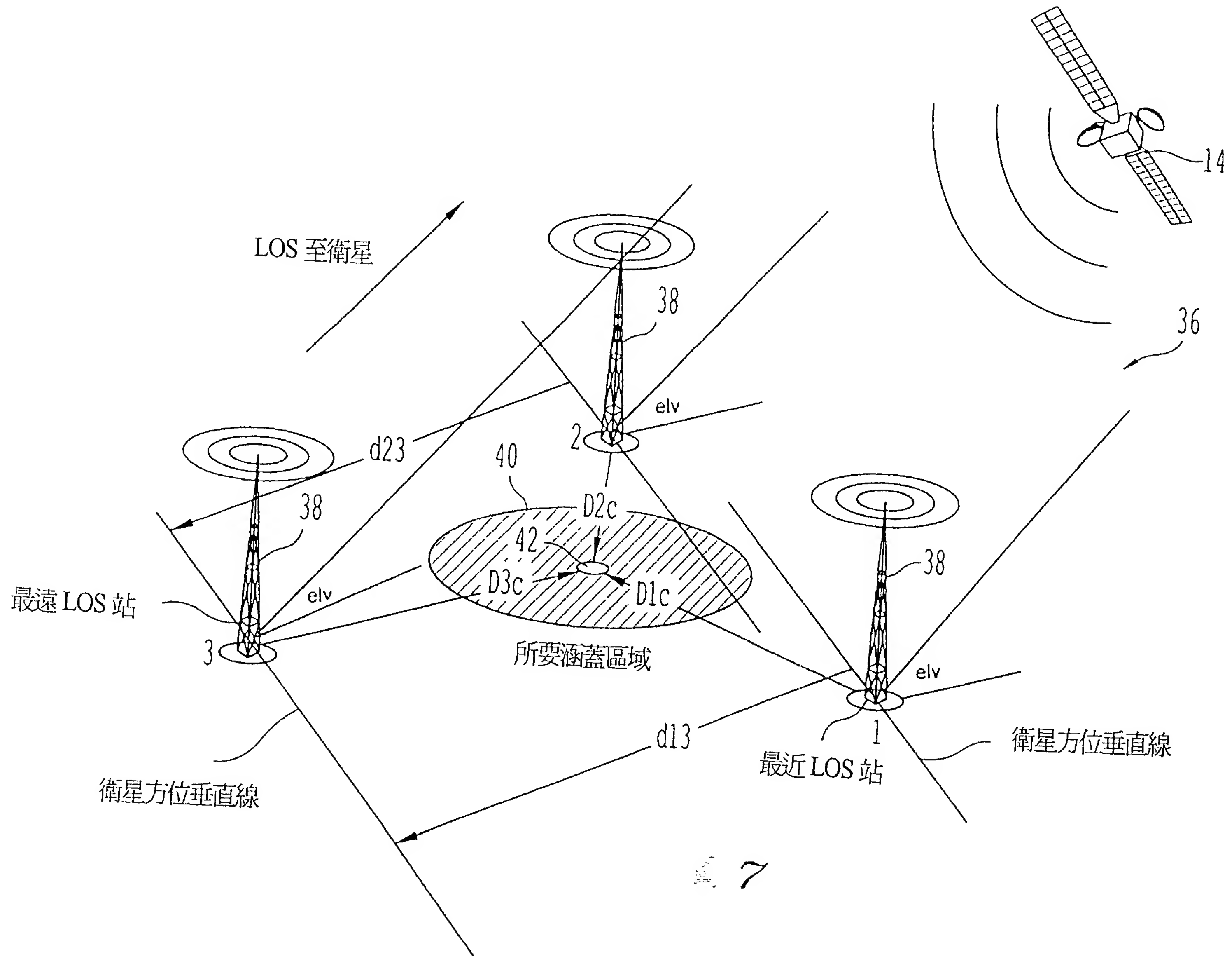
每一符號 2 位元之 TDM
流及符號速率 R
位元速率=2 x R
52
N_{TDM}
符號
54
N_{TDM}
係數
56
護衛期間時間壓縮
58
MCM 框時間壓縮
MCM 框長度 = M_{MCM} +Sync
MCM 符號
同步字元
護衛期間

圖 6



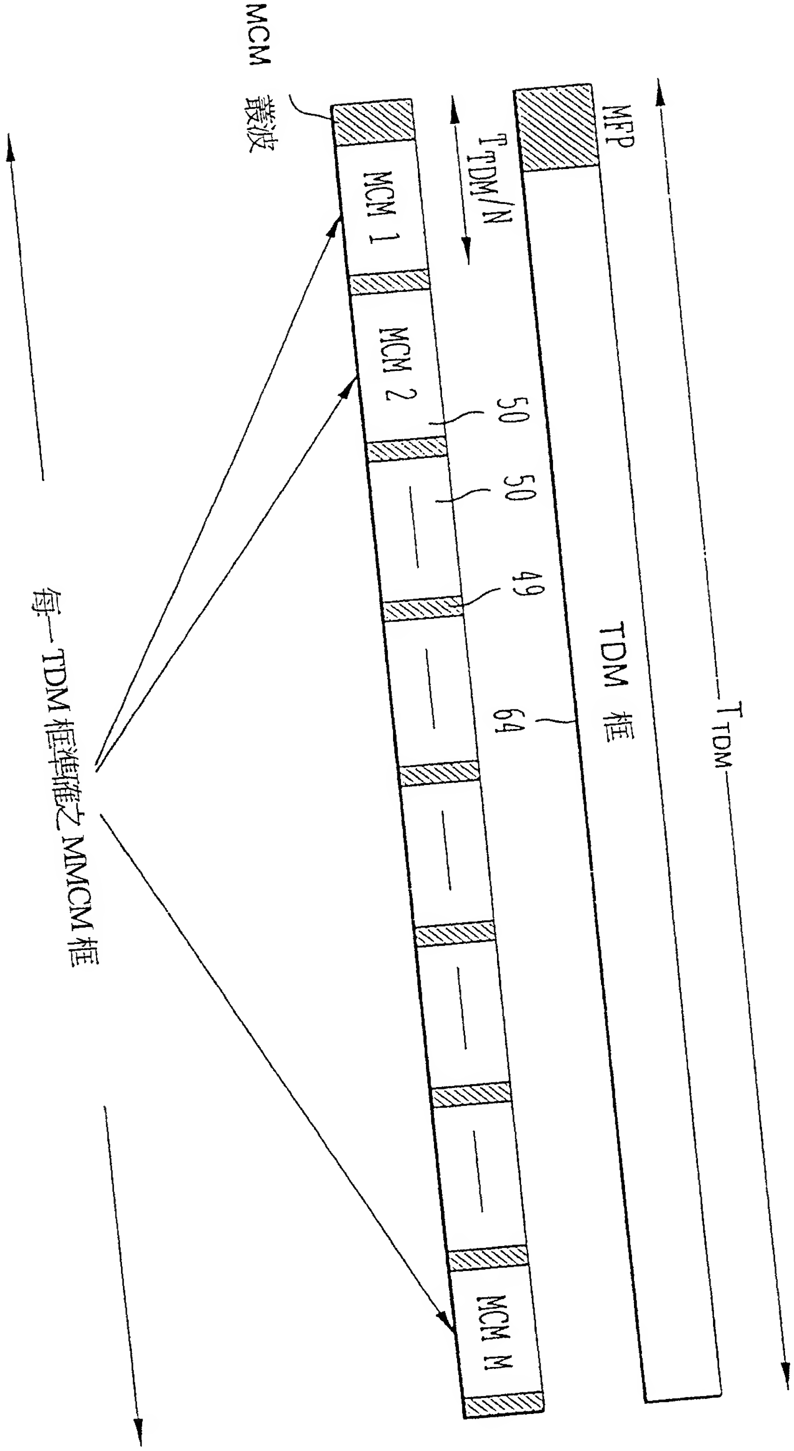
衛星視線
傾斜範圍 = $dk_M \times \cos elv$
elv
m
k
dk_M
水平面距離

圖 8



14
36
LOS 至衛星
38
38
38
elv
elv
elv
2
3
1
d23
d13
40
42
D2c
D3c
D1c
所要涵蓋區域
最遠 LOS 站
最近 LOS 站
衛星方位垂直線
衛星方位垂直線

圖 7


MFP
T_{TDM}
TDM 框
$T_{TDM/N}$
50
50
49
64
MCM 載波
MCM 1
MCM 2
MCM M
每一TDM框準確之MMCM框

9

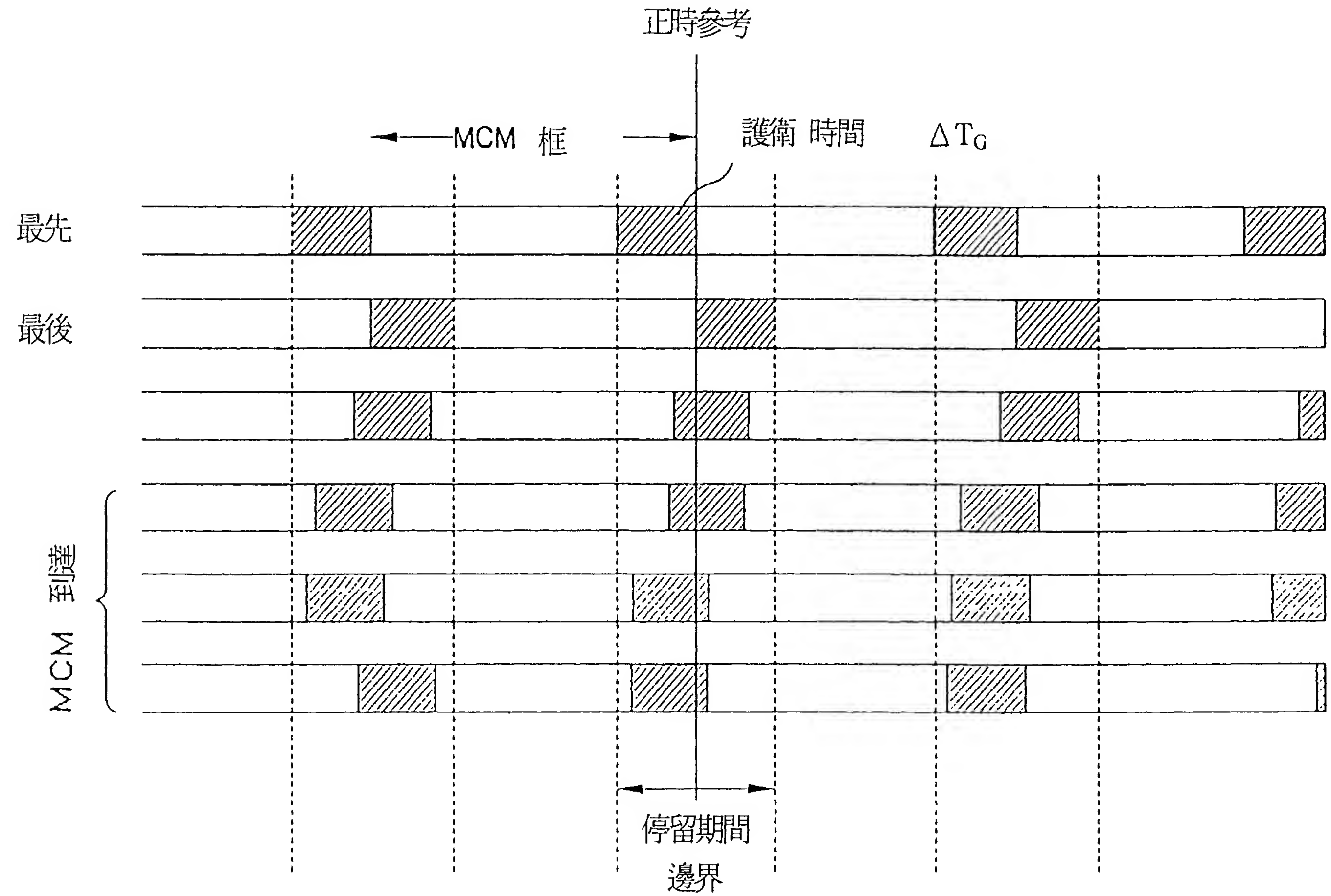


圖10

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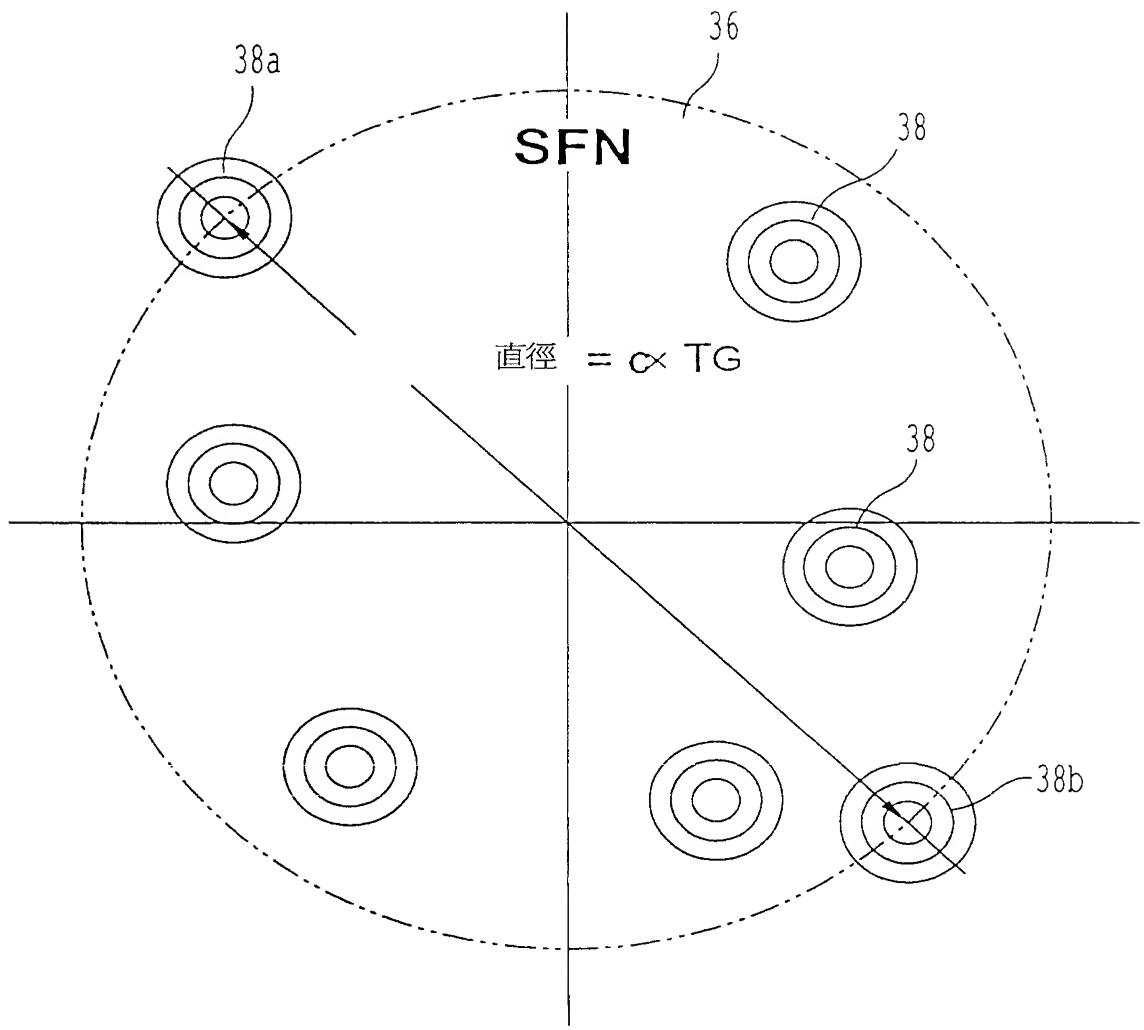

圖 11

Coding of data stream

Patent/Publication Number	583842
Issued/Publication Date	2004/04/11
Application Date	2001/09/06
Application Number	090122097
Certification Number	199477
IPC	H03M-013/35
Inventor	MARTINI, MARIA GIUSEPPINA IT; CHIANI, MARCO IT
Applicant	KONINKLIJKE PHILIPS ELECTRONICS N.V.NL
Priority Number	20000717 EP 20000202531

Abstract

Coding a data stream is provided, wherein the data stream comprises at least one packet having a given packet length and respective partitions of the at least one packet are coded with different error protection rates, the respective lengths of the respective partitions being determined by respective predetermined percentages of the packet length or a fraction of the packet length.

申請日期	90.9.6
案號	90122097
類別	H03M 13/35

(以上各欄由本局填註)

A4
C4

583842

~~新型~~ 發明 專利說明書

一、~~新型~~ 發明 名稱	中文	資料流之編碼
	英文	CODING OF DATA STREAM
二、~~創作~~ 發明 人	姓名	1. 瑪利亞 吉賽皮娜 馬丁尼 MARIA GIUSEPPINA MARTINI 2. 瑪可 奇亞尼 MARCO CHIANI
	國籍	均義大利
	住、居所	1.義大利培魯吉亞市皮亞薩單提路7號 2.義大利利米尼市賈拉伯布路24號
三、申請人	姓名(名稱)	荷蘭商皇家飛利浦電子股份有限公司 KONINKLIJKE PHILIPS ELECTRONICS N. V.
	國籍	荷蘭
	住、居所(事務所)	荷蘭愛因和文市格羅尼渥街1號
	代表人姓名	J. L. 凡 德 渥 J. L. VAN DER VEER

裝 訂 線

經濟部智慧財產局員工消費合作社印製

(由本局填寫)

承辦人代碼：	
大　　類：	
ＩＰＣ分類：	

A6
B6

本案已向：

國（地區）	申請專利，申請日期：	案號：	，□有　□無主張優先權
歐洲專利機構	2000年07月17日	00202531.0	□有　☑無主張優先權

有關微生物已寄存於：　　　　　　　　　　，寄存日期：　　　　　　　　　　，寄存號碼：

(請先閱讀背面之注意事項再填寫本頁各欄)

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經濟部智慧財產局員工消費合作社印製

四、中文發明摘要（發明之名稱：　資料流之編碼　）

本發明揭示一種資料流之編碼，其中資料流包括至少一具有一給定封包長度的封包，並且該至少一封包的各自分割部份係使用不同的誤差保護率編碼，該等各自分割部份的各自長度係按封包長度之各自預先決定百分比決定，或按該封包長度之分數決定。

英文發明摘要（發明之名稱：CODING OF DATA STREAM　）

Coding a data stream is provided, wherein the data stream comprises at least one packet having a given packet length and respective partitions of the at least one packet are coded with different error protection rates, the respective lengths of the respective partitions being determined by respective predetermined percentages of the packet length or a fraction of the packet length.

（請先閱讀背面之注意事項再填寫本頁各欄）

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五、發明說明（ 1 ）

發明領域

本發明與資料流之編碼及解碼有關。

本發明進一步與資料流之傳輸及接收有關。

發明背景

請參考 2000 年 1 月 IEEE Signal Processing Magazine 中 M. Budagavi、W. Rabiner Heinzelman、J. Webb、R. Talluri 著作的"Wireless MPEG-4 Video Communication on DSP Chips"文章。該篇文章發表，為了使壓縮位元流更健全，MPEG-4 視訊壓縮標準將數個錯誤復原工具併入它的簡單分析工具 (simple profile) 中，以能夠偵測、包含及隱藏錯誤。這些是功能強大的來源編碼技術，用以對付位元錯誤小於 10^{-3} 的情況；但是，現今無線通道可具有更高的位元錯誤率 (BER)。有關行動無線通道的惡劣狀況係起因於在發射器與接收器之間移動而導致的多路徑衰落，以及地面環境的變化。多路徑衰落本身顯然是長錯誤叢發的形式。因此，需要某種交錯及通道編碼形成，以改良通道狀況。使用來源編碼與通道編碼的組合，可運用 MPEG-4 簡單分析 (simple-profile) 視訊壓縮，在易錯誤 (error-prone) 無線通道上實現可接受的視覺品質。MPEG-4 壓縮位元流結構也適合使用非相等錯誤保護 (unequal error protection；UEP) (一種結合來源-通道編碼的形式)，以確定位元流重要部份的錯誤較少。

發明概要

本發明的目的是提供一種改良的資料傳輸。為此目的，

五、發明說明（ 2 ）

本發明提供如獨立申請專利範圍中定義的編碼、解碼、傳輸、接收、資料流及儲存媒體。在依附的申請專利範圍中定義有利的具體實施例。

本發明特別適用於 MPEG-4 視訊無線傳輸領域。本發明人認識到 MPEG-4 封包的長度不是完全一樣長，並且不同封包中的分割部份具有不同的長度，這是因為使用可變長度編碼，以及每個封包中具有整數巨集區塊數量的需求。這意謂著無法使用 UEP 機制，並且為了運用正確編碼率來執行解碼，接收器應知道通道解碼層級的位元流結構。封包（似分割部份）的長度不同；因此，應針對每個封包來動態變更 UEP 機制，並且必須知道分割部份長度。本發明提供適用於具有可變長度之封包及分割部份的 UEP。

根據本發明第一項觀點，資料流中至少一封包的各自分割部份係使用不同的誤差保護率編碼，其中該等各自分割部份的各自長度係按該至少一封包長度之各自預先決定百分比決定，或按該封包長度之分數決定。藉由依據固定百分比提供分割，可實行適用於具有可變長度的 UEP。

在實用的具體實施例中，整個封包內所有分割部份的長度係按該封包長度的百分比決定。但是，某些分割部份的長度也可採用固定、預先決定長度來決定。然後，剩餘分割部份的長度最好是按該封包長度的分數百分比決定。這個分數通常等於該等分割部份總長度，該等分割部份的長度係按（分數的）百分比決定。在實用具體實施例中，這等於封包長度減固定長度的總和。所以，可組合固定分割部

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五、發明說明（3）

份長度及比例分割部份長度。

其優點為，給定的封包長度係按照介於資料流中兩個標記(marker)間的距離決定，其中該等兩個標記的至少一標記指示封包開始。

其優點為，其中所決定的各自預先決定百分比促使該封包的一第一分割部份包含至少一第一原始封包分割部份。該第一原始封包分割部份可能是該封包的標題。藉由在正常條件下選取第一百分比，使該標題一定包含於該第一分割部份中，可使用一相同保護率來保護整個標題，該保護率最好高於後續分割部份的保護率。進一步決定的百分比最好促使給定分割部份與前一分割部份的總和最好與原始分割部份數量相同。

在根據本發明一項具體實施例的解碼器中，一資料流包括至少一具有一給定封包長度的封包，其中該至少一封包的各自分割部份係使用不同的誤差保護率編碼，接收，以及該等各自分割部份係使用不同的誤差保護率解碼。

參考下文中詳細說明的具體實施例即可明白本發明的前述及其他觀點。

圖式簡單說明

圖 1 顯示 MPEG-4 中的資料分割；

圖 2 顯示根據本發明一項具體實施例之保護機制的原理圖；

圖 3 顯示根據本發明一項具體實施例之開始碼代換(start code substitution)及(成比例)非相等錯誤保護；

五、發明說明（4　）

圖 4 顯示本發明第一項具體實施例之發射器，該發射器包括開始碼偵測及代換裝置；

圖 5 顯示本發明第一項具體實施例之接收器，該接收器包括開始碼偵測及更換裝置；

圖 6 顯示根據本發明一項具體實施例之成比例非相等錯誤保護。

圖式中只顯示瞭解本發明所需的元件。

圖式詳細說明

由於壓縮，尤其使用預測編碼及可變長度編碼（Variable Length Coding；VLC），導致 MPEG-4 位元流非常易受到錯誤的影響。1988 年 6 月 IEEE Communication Magazine 第 36 卷第 6 號，R. Talluri 著作的"Error-resilient video coding in the ISO MPEG-4 standard"文章中說明於 ISO MPEG-4 標準中標準化的視訊編碼技術的錯誤復原觀點。ISO MPEG-4 標準中採納特定工具，使透過有雜訊之無線通道上傳達的壓縮視訊資料能夠詳細呈現。這些技術包括重新同步化策略、資料分割、可反相可變長度編碼（Variable Length Code）及標題延伸編碼。

這些工具有且於加強 MPEG-4 位元流的強固性。配合使用重新同步化（Resync）標記，MPEG-4 位元流結果由長度幾乎相同的封包所組成。不管此類的工具，當透過無線通道傳輸 MPEG-4 時，可實現的接收品質仍然不佳。但是，如果在通道編碼層級利用錯誤復原工具，則錯誤復原工具可進一步改良所接收的視訊品質。具體而言，可有效

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五、發明說明（ 5 ）

利用資料分割工具來執行非相等錯誤保護（unequal error protection；UEP）：將包含於每個封包中的資訊位元分割成三個分割部份，每個分割部份均具有不同的通道錯誤靈敏度。如圖 1 所示，針對 I 訊框，分割部份係由標題 HI 與藉由 DC 標記 DCm 所分隔的 DC DCT 係數和 AC 係數所組成。針對連接的 P 訊框，分割部份係由標題 HP 與藉由運動標記 mm 所分隔的運動分割部份和結構分割部份 tp 所組成。

一種適用的技術同時考慮無線通道及所說明之應用的特性。具體而言，應透過 UEP 來利用關於來源位元對通道錯誤的不同靈敏度。這個技術在於依據來源位元對通道錯誤的感知靈敏度來執行錯誤保護：愈敏感的位元使用愈高的保護（對應於較低保護率編碼）來保護，針對低重要性的位元則使用較低的保護（即，較高保護率編碼）。與典型的正向錯誤校正（Forward Error Correction；FEC）相比，UEP 允許透過利用來源的特性，以相同的位元速率來實現較高的感知視訊品質。

在建議的機制中，會依據相關資訊的主觀重要性，使用不同的編碼率來保護這三個分割部份。內含於標題的資訊對連續解碼封包而言非常重要，因此應極力保護這些資訊。就訊框內而言，DC 係數的主觀重要性高於 AC 係數；因此，對 DC 係數的保護應高於 AC 係數。就感知訊框而言，運動資料受到的保護應高於結構資料，就如同運動資料被正確接收，而結構資料可能會被局部重新建構。

五、發明說明(6)

建議的 UEP 實施也考慮不同類型訊框的不同重要性:在 MPEG-4 標準中。內部(Intra)、預測(Predicted) 及回溯預測 (Backward predicted) 訊框均被考慮,其中內部(Intra) 訊框係以獨立方向編碼,而與其他無關,而預測(Predicted) 訊框利用來自於連續訊框的資訊。

正確接收內部(Intra)訊框對執行後續預測(Predicted) 訊框的運動補償而言非常重要,因此較低平均通道編碼率(即,較高保護)應與內部(Intra)訊框組合在一起,而預測(Predicted)訊框可使用較高平均通道編碼率(即,較低保護)編碼。圖 2 顯示所期望之保護機制的原理表示。

UEP 可能係透過速率相容擊穿捲積(Rate Compatible Punctured Convolutional;RCPC)編碼執行,其中速率係依據位元的感知重要性而定。在此情況下,考慮的編碼係藉由擊穿 (puncturing) 同一"母"編碼的方式獲得。然後,只需要一個編碼器及一個解碼器就可執行整個位元流的編碼及解碼。速率相容擊穿捲積編碼 (Rate Compatible Punctured Convolutional Code)本身可從 1988 年 4 月 IEEE Trans. Commun.,第 36 卷第 4 號第 389 至 400 頁 J. Hagenauer 著作的 "Rate-Compatible Punctured Convolutional Codes (RCPC Codes) and their Applications"文章中得知。

考慮不同的平均編碼率來保護不同的訊框 (I 訊框係使用較高保護/較低速率編碼,P 訊框則考慮使用較低保護/較高速率編碼),並且針對每個訊框,利用加入至 MPEG-4 標

準的資料分割工具，以為最重要的分割部份提供更強的保護。如果未正確接收訊框，則可重新傳輸訊框。

MPEG-4 編碼位元流的結構為：視訊物件 (Video Object；VO)、視訊物件層 (Video Object Layer；VOL)、視訊物件平面群組 (Groups of Video Object Plane；GOV)、視訊物件平面 (Video Object Plane；VOP) 及封包。為了允許同步化，會藉由相關的開始碼來標示位元流每個部份的開始。開始碼是唯一的字組，可從任何合法的可變長度編碼字組序列辨識。H1 標示 VO 的開始碼、H2 標示 VOL 的開始碼、H3 標示 GOV 的開始碼、H4 標示 VOP 的開始碼及 H5 標示封包開始碼 (重新同步化 (Resync))。

主要的問題是 MPEG-4 封包的長度不是完全一樣長，並且不同封包中的分割部份具有不同的長度，這是因為使用可變長度編碼，以及每個封包中具有整數巨集區塊數量的需求。這意謂著無法使用 UEP 機制，並且為了運用正確編碼率來執行解碼，接收器應知道通道解碼層級的位元流結構。封包 (似分割部份) 的長度不同；因此，應針對每個封包來動態變更 UEP 機制，並且必須知道分割部份長度。由於考慮到這個問題，因而建議一種用以執行 UEP 的解決方案：成比例 UEP (Proportional UEP)。

成比例非相等錯誤保護 (Proportional Unequal Error Protection；P-UEP)

圖 6 顯示成比例非相等錯誤保護 (Proportional

五、發明說明（ 8 ）

Unequal Error Protection）的原理。由於接收器不知道每個欄位的長度，所以使用一種成比例機制，給定封包的（可變）長度。最好透過接收兩個適當的開始碼（至少其中一個是封包開始碼）來決定封包長度。為了填寫封包緩衝器，此類的機制會造成一個封包延遲。考慮位元流的特性，為每個分割部份選取百分比長度。給定三個分割部份的百分比長度為 P_1、P_2、P_3，受到速度 R_1、R_2、R_3 保護，則給定 I 封包平均速度的方式為：

$$R_{avg}=\frac{R_1R_2R_3}{P_1R_2R_3+P_2R_1R_3+P_3R_1R_2}$$

同樣地，針對 P 封包：

$$R_{avg}'=\frac{R_1'R_2'R_3'}{P_1R_2'R_3'+P_2R_1'R_3'+P_3R_1'R_2'}$$

接著，已編碼封包的長度為：

$$L_{coded\_packet\_I}=\frac{L_{packet}}{R_{avg}}+\frac{M}{R_3}$$

適用於 I 訊框

以及

$$L_{coded\_packet\_p}=\frac{L_{packet}}{R_{avg}'}+\frac{M}{R_3'}$$

適用於 I 訊框

其中 M 是編碼的記憶體，在此情況下考慮捲積編碼。就編碼的記憶體 M 而言：捲積編碼與區塊編碼的差異為，編碼器包含記憶體，並且在任何給定時間單元，編碼器輸出不僅取決定該時間單元的輸入，而且還取決於前 M 個輸入區塊，其中 M 是編碼的記憶體。記憶體 M 捲積編碼器係由 M 階移位暫存器所組成，其中所選階的輸出經過模 2 (modulo-2) 相加，以構成已編碼符號。由於捲積編碼器屬

五、發明說明（ 9 ）

於連續電路，其運作可藉由狀態圖來說明。編碼器的狀態被定義成它的移位暫存器內容；因此編碼器可假設 2^M 個狀態。為了使用相同的位元強度來保護位元流最後的位元，應將 M 個尾端位元加入至位元流，以強制編碼器叢集回到已知狀態（通常是"0"狀態）。事實上，如果考慮捲積編碼，則終止封包的方式為將 M 位"0"位元移位至移位暫存器，以便允許適當終止格子。尾端位元係使用較高速率編碼。為了計算總平均速度，應計算介於 I 訊框與 P 訊框之間的平均值，並且也應考慮因開始碼代換所造成的內部操作(overhead)。

本發明的觀點採用各別預先決定的可變封包長度百分比當作各別的封包分割部份。考慮資料流的特性，所決定的百分比最好促使該封包的第一分割部份包含至少第一原始封包分割部份（例如，標題），並且第一分割部份與第二分割部份的總和包含至少第一原始封包分割部份及第二原始封包分割部份等等。

第二項主要問題為 MPEG-4 開始碼不穩定而容易錯誤：開始碼中的單一錯誤會造成偵測失誤，導致同步化損失。為了應付這些問題，本發明建議一些有利的解決方案。如果發生錯誤，可模擬開始碼以及偵測失誤。為了解決這個問題，建議開始碼代換。

開始碼代換

在進一步具體實施例中，在 MPEG-4 編碼（請參閱圖 3）之後，會使用偽雜訊字組來替換開始碼，其中偽雜訊字組

五、發明說明（ 10 ）

是具有高關聯性屬性的序列，例如 Gold 序列）。這個新開始碼係以無線開始碼（Wireless Start Code）標示。具體而言，針對 VO、VOL、VOP、GOV 開始碼及重新同步化（Resync）標記執行代換。圖 3 顯示已編碼資料流 S，其包含標記 H1...H5。使用具有較高防通道錯誤之強固性的標記 WH1...WH5 來代換這些標記，以獲得適用於無線傳輸的資料流 WS。接收器接收到的資料流 WS 變成資料流 RS，其類似於 WS 但可能有通道錯誤。標記 WH1...WH5 被接收而成為 $WH1_R...WH5_R$。標記（字組）$WH1_R ...WH5_R$ 類似於 WH1...WH5，但可能有通道錯誤。因為這些標記具有較高的關聯性屬性，所以可被辨識為 WH1...WH5，之後分別被類似於 H1 ...H5 的標記代換。就 MPEG-4 位元流而論，圖 3 中的資料流不包含 GOV 開始碼（H3）。在 MPEG-4 位元流中，在 VOL 開始碼（H2）之後沒有 GOV 開始碼（H3），這是因為 VOL 開始碼（H2）也包含 GOV 的開頭。

於接收器端，在通道解碼處理程序之前，會先透過關聯性來評估這些無線開始碼 WH1...WH5 的位置；介於可能缺少開始碼與可能評估開始碼之間應達成交換，以此方式相應選擇無線開始碼長度及關聯性的適當門限值。隨著執行偵測，使用來自於原始開始碼集的對應開始碼 H1...H5 來代換無線開始碼 WH1...WH5。藉此使 MPEG-4 解碼器（請參閱圖 5）完全明白所要的代換。

在通道編碼層級，建議根據本發明的優秀具體實施例：

五、發明說明（ 11 ）

開始碼代換 (Start Codes Substitution) 結合成比例非相等錯誤保護 (Proportional Unequal Error Protection；P-UEP)。

就相符訊框的 VOP 簡化案例，提供優秀具體實施例的說明。圖 4 與 5 中的虛線標示控制線。

圖 4 顯示根據本發明的發射器，該發射器包括開始碼偵測器 12，用以偵測開始碼 H1...H5。偽雜訊字組產生器 13 使用對應的偽雜訊字組 WH1...WH5 來代換偵測到的開始碼。將偽雜訊字組 WH1...WH5 提供給多工器 14，其包括要傳輸之資料流 WS 中的偽雜訊字組。

於封包緩衝器 10 中接收資料流 S。在通道編碼器 11 對出現在標記 H1...H5 之間的資料流 S 封包進行通道編碼，以獲得通道編碼封包。將這些通道編碼封包提供給多工器，並且納入要傳輸的資料流 WS 中。將要傳輸的資料流提供給天線（例如，進行無線傳輸）或提供給儲存媒體 15。

如上文所述，使用 P-UEP 執行圖 4 所示的通道編碼具有許多優點，但是也可使用其他的通道編碼機制。

圖 5 顯示接收器 3，用以接收圖 4 所示之發射器所傳輸的資料流 WS。在開始碼偵測器 32（例如，偽雜訊字組偵測器）中，執行介於每個允許的偽雜訊字組（即，來自於預先決定偽雜訊字組集，對應於標記）與相關位元流部份之間的關聯性評估，以偵測代表開始碼的偽雜訊字組。關聯性係比對對應的門限值 th。當偵測偽雜訊字組時，位元流中

的位元指示項移位適當數量的位元，並由開始碼產生器 33 提供對應的 MPEG- 4 開始碼 H1...H5，將開始碼插入至多工器 34 中，由多工器負責排列要饋送至 MPEG-4 解碼的位元流 S。如果偵測到 GOV 開始碼或 VOP 開始碼，則 VOP 指示項會變更其狀態。

如果偵測到重新同步化 (Resync) 標記，則會初始化封包緩衝器 30，並將後續位元填入緩衝器，直到偵測到下一個開始碼。除非緩衝器包含 N 位位元，否則不會執行關聯性評估，其中 N 是封包的最小長度。當偵測到下一個開始碼時，緩衝器 30 內含一個封包；依據 VOP 指示項 infn 及百分比，在通道解碼器 31 中十對緩衝器中的位元執行通道解碼。這個結構中使用的速率最好固定，並且與通道編碼器 11 中使手的速度相同。就可變速率而言，發射器 1 必須接收來自於通道編碼器 11 的速率。將經過通道解碼的封包插入至多工器 34 中，由多工器負責排列要饋送至 MPEG-4 解碼的位元流。請注意，如果使用 RCPC 編碼，則會在解碼之前執行解擊穿(de-puncturing)。在此情況下，然後以母編碼速率 (mother code rate) 將封包解碼。

雖然圖 4 和 5 中未顯示，但是發射器中的調變器可先將資料流調變，之後才傳輸資料流，因此，在執行解碼之前，先在接收器中的解調變器將已調變的資料流解調變。

請注意，上述的具體實施例是用於說明本發明，而不是用於限定本發明，熟知技藝人士能夠設計許多替代具體實施例，而不會脫離隨附申請專利範圍的範疇。在申請專利

五、發明說明（ 13 ）

範圍中，放置在圓括號內的任何參照符號不應視為限制該項申請專利範圍。申請專利範圍中的"包括"並不排除使用其他的元件或步驟。本發明可藉由包含數個不同元件的硬體實施，或藉由經過適當程式規劃的電腦實現。在裝置申請專利範圍中列舉的數個裝置中，可用一個及相同硬體項目將這些裝置具體化。這僅僅是在互相不同的相依申請專利範圍中列舉特定措施，而不是表示無法有效這些措施的組合。

簡言之，本發明揭示一種資料流之編碼，其中資料流包括至少一具有一給定封包長度的封包，並且該至少一封包的各自分割部份係使用不同的誤差保護率編碼，該等各自分割部份的各自長度係按封包長度之各自預先決定百分比決定，或按該封包長度之分數決定。

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六、申請專利範圍

1. 一種資料流之編碼方法，該資料流包括至少一具有一給定封包長度的封包，該方法包括下列步驟：

   使用不同的誤差保護率編碼該至少一封包的各自分割部份，其中該等各自分割部份的各自長度係按該封包長度之各自預先決定百分比決定，或按該封包長度之分數決定；以及

   輸出具有以該等不同的誤差保護率編碼之該至少一封包之各自分割部份的該資料流。

2. 如申請專利範圍第 1 項之方法，其中該給定的封包長度係按照介於該資料流中之兩個標記 (marker) 間的距離決定，其中該等兩個標記的至少一標記指示一封包開始。

3. 如申請專利範圍第 1 項之方法，其中所決定的各自預先決定百分比促使該封包的一第一分割部份包含至少一第一原始封包分割部份。

4. 如申請專利範圍第 3 項之方法，其中所決定的各自預先決定百分比促使該封包的該第一分割部份與一第二分割部份的總和包含至少該第一原始封包分割部份及一第二原始封包分割部份。

5. 一種資料流之編碼方法，接收到的資料流包括至少一具有一給定封包長度的封包，其中該至少一封包的各自分割部份已使用不同的誤差保護率編碼，該等各自分割部份的各自長度係按該封包長度之各自預先決定百分比決定，該方法包括下列步驟：

   使用該等不同的錯誤保護率將該等各自封包分割部份解

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六、申請專利範圍

碼；以及

輸出具有以該等不同的誤差保護率解碼之各自分割部份的該資料流。

6. 一種編碼一資料流之編碼器，該資料流包括至少一具有一給定封包長度的封包，該編碼器包括：

裝置，用以使用不同的誤差保護率編碼該至少一封包的各自分割部份，其中該等各自分割部份的各自長度係按該封包長度之各自預先決定百分比決定，或按該封包長度之分數決定；以及

裝置，用以輸出具有以該等不同的誤差保護率編碼之該至少一封包之各自分割部份的該資料流。

7. 一種解碼一資料流之解碼器，接收到的資料流包括至少一具有一給定封包長度的封包，其中該至少一封包的各自分割部份已使用不同的誤差保護率編碼，該等各自分割部份的各自長度係按該封包長度之各自預先決定百分比決定，或按該封包長度之分數決定，該解碼器包括：

裝置，用以使用該等不同的錯誤保護率將該等各自封包分割部份解碼；以及

裝置，用以輸出具有以該等不同的誤差保護率解碼之各自分割部份的該資料流。

8. 一種用來傳輸一資料流之發射器，該發射器包括：

如申請專利範圍第 6 項之編碼器；以及

天線裝置，用以傳輸該資料流。

9. 一種用來接收一資料流之接收器，該接收器包括：

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六、申請專利範圍

天線裝置，用以接收該資料流；以及
如申請專利範圍第 7 項之解碼器。

10. 一種資料流，該資料流包括至少一具有一給定封包長度的封包，其中該至少一封包的各自分割部份已使用不同的誤差保護率編碼，該等各自分割部份的各自長度係按該封包長度之各自預先決定百分比決定，或按該封包長度之分數決定。

11. 一種已儲存資料流之儲存媒體，其中該資料流包括至少一具有一給定封包長度的封包，其中該至少一封包的各自分割部份已使用不同的誤差保護率編碼，該等各自分割部份的各自長度係按該封包長度之各自預先決定百分比決定，或按該封包長度之分數決定。

裝

訂

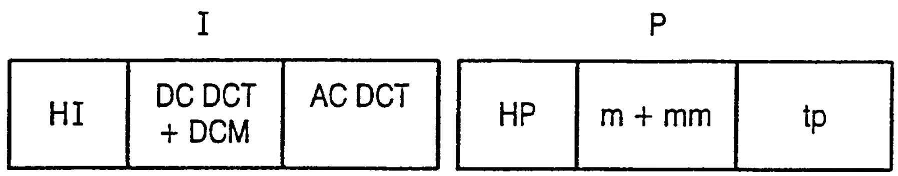
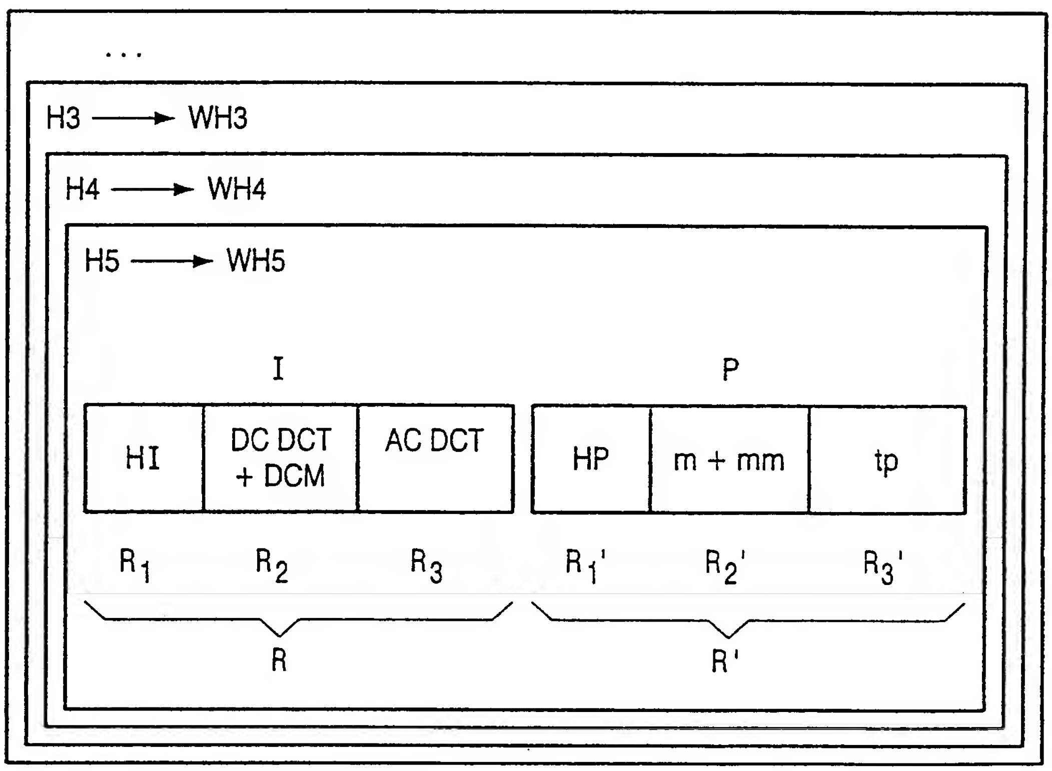

I
P
HI
DC DCT
+ DCM
AC DCT
HP
m + mm
tp

圖 1


...
H3 → WH3
H4 → WH4
H5 → WH5
I
P
HI
DC DCT
+ DCM
AC DCT
HP
m + mm
tp
R1
R2
R3
R1'
R2'
R3'
R
R'

R < R'

圖 2

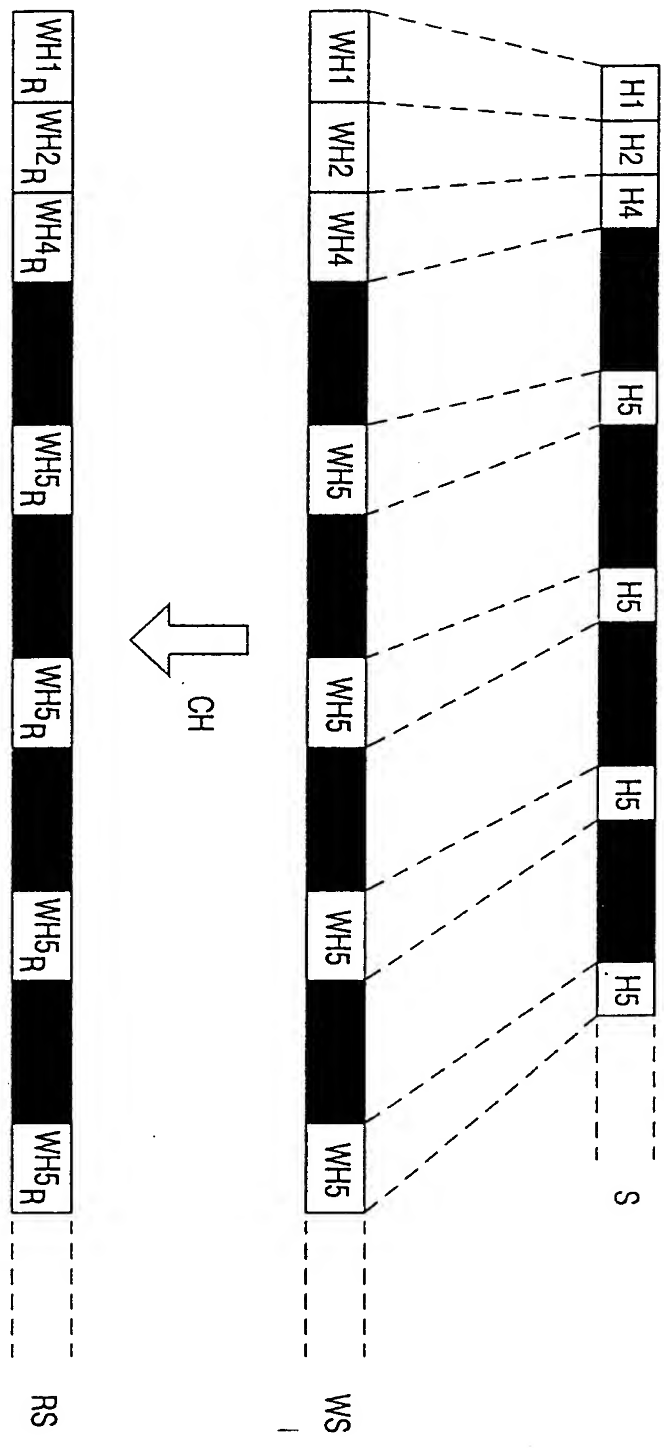

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圖 3

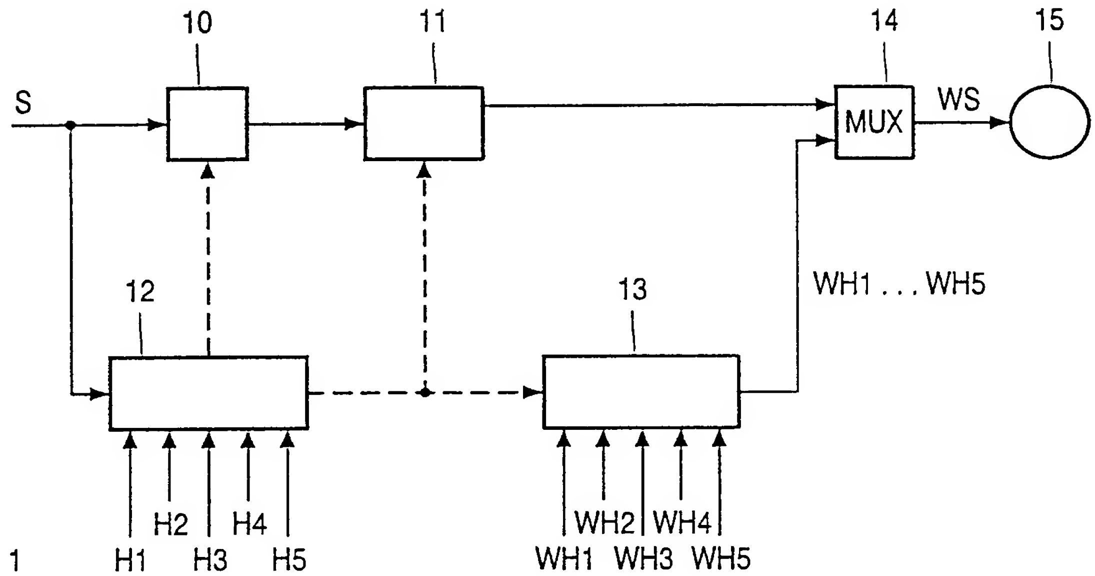
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圖 4

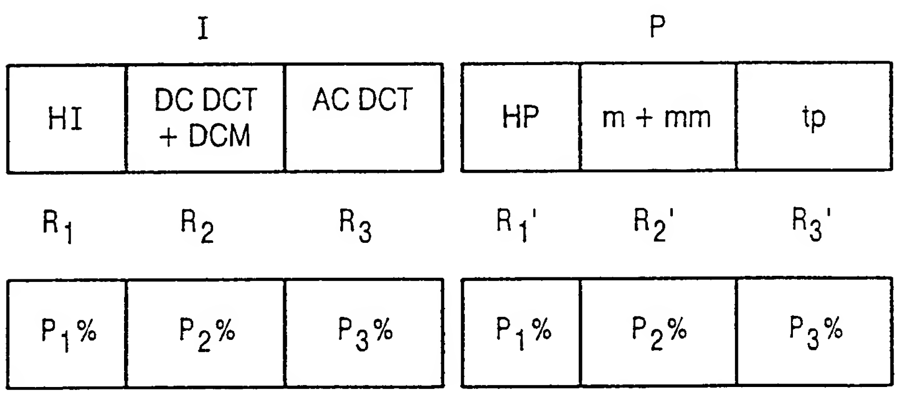
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圖 6

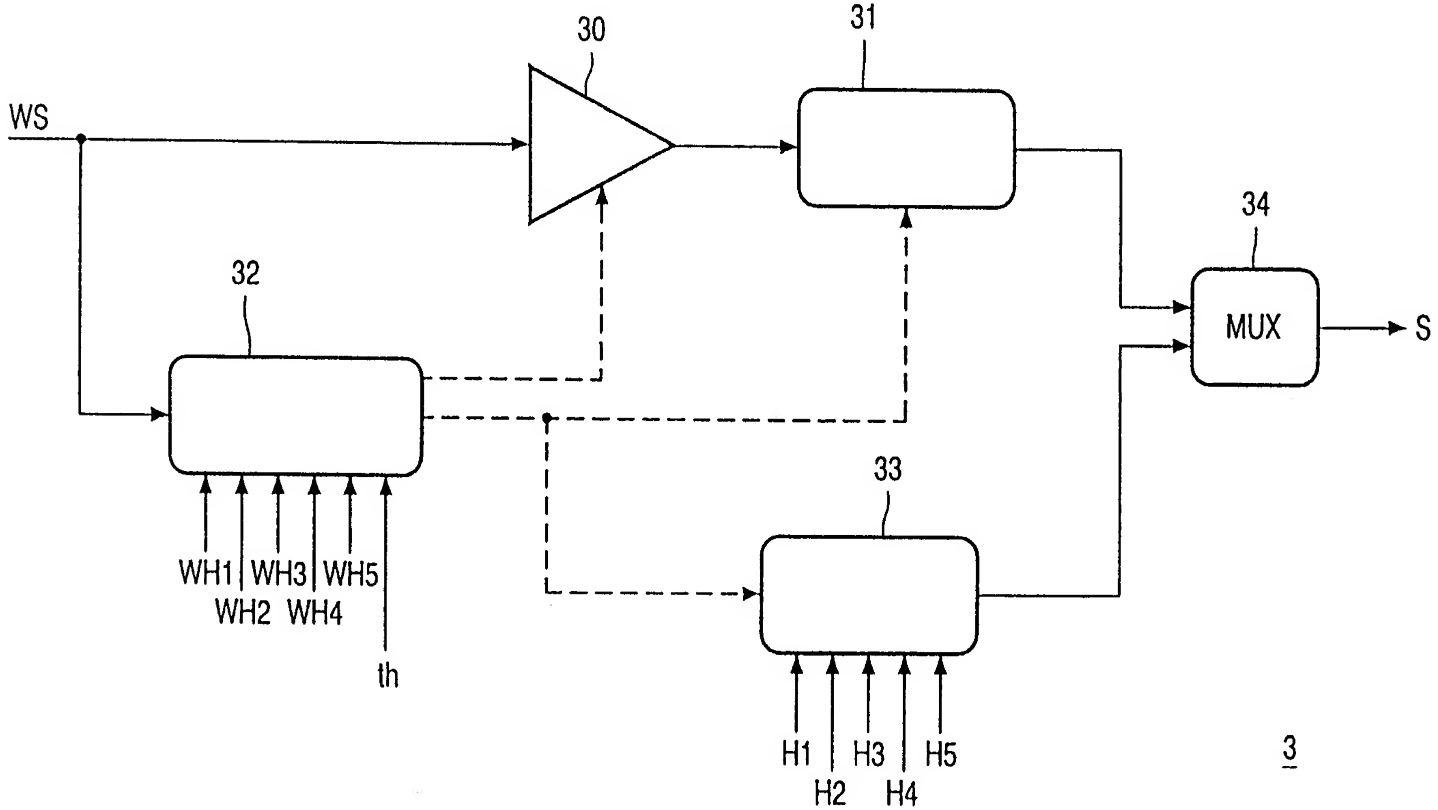

圖 5

92.11.13 修正

五、發明說明 (14)

圖式元件符號說明

1	發射器
3	接收器
10	封包緩衝器
11	通道編碼器
12	開始碼偵測器
13	偽雜訊字組產生器
14	多工器
15	儲存媒體
30	封包緩衝器
31	通道解碼器
32	開始碼偵測器
33	開始碼產生器
34	多工器

裝 訂 線

(12) DEMANDE INTERNATIONALE PUBLIÉE EN VERTU DU TRAITÉ DE COOPÉRATION EN MATIÈRE DE BREVETS (PCT)

(19) Organisation Mondiale de la Propriété Intellectuelle
Bureau international

(43) Date de la publication internationale
6 février 2003 (06.02.2003)

PCT

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WO 03/010994 A1

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(71) Déposant *(pour tous les États désignés sauf US)* **: NEXO** [FR/FR]; 154, rue des Erables, ZAC de Paris Nord II, F-93420 Villepinte (FR).

(72) Inventeur; et
(75) Inventeur/Déposant *(pour US seulement)* **: VINCENOT, Eric** [FR/FR]; 1 à 5, rue Pixerecourt, F-75020 Paris (FR).

(74) Mandataire : CABINET BONNET-THIRION; 12, avenue de la Grande Armée, Boîte postale 966, F-75829 Paris Cedex 17 (FR).

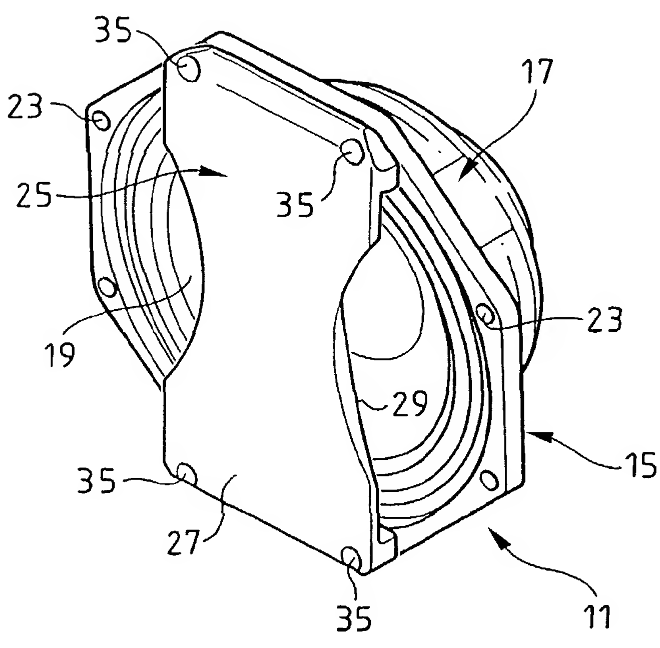
(81) États désignés *(national)* **:** AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, OM, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VN, YU, ZA, ZM, ZW.

(84) États désignés *(régional)* **:** brevet ARIPO (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), brevet eurasien (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), brevet européen (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, SK, TR), brevet OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

[Suite sur la page suivante]

(54) Title: LOUDSPEAKER WITH DIRECT EMISSION AND OPTIMISED RADIATION

(54) Titre : HAUT-PARLEUR A RADIATION DIRECTE ET RAYONNEMENT OPTIMISE



(57) Abstract: The invention concerns a mobile membrane loudspeaker equipped with a partial closure optimising its radiation. The invention is characterised in that the loudspeaker with mobile diaphragm (19) attached to a rigid frame (15) defining an acoustic emission plane (P) comprises a closure (25) for only one central zone of said emission plane.

(57) Abrégé : Haut-parleur à membrane mobile équipé d'un obturateur partiel optimisant son rayonnement. Selon l'invention, le haut-parleur à membrane mobile (19) rattachée à un châssis rigide (15) définissant un plan d'émission acoustique (P) comporte un obturateur (25) de seulement une zone centrale dudit plan d'émission.

WO 03/010994 A1

Publiée :

— *avec rapport de recherche internationale*
— *avant l'expiration du délai prévu pour la modification des revendications, sera republiée si des modifications sont reçues*

En ce qui concerne les codes à deux lettres et autres abréviations, se référer aux "Notes explicatives relatives aux codes et abréviations" figurant au début de chaque numéro ordinaire de la Gazette du PCT.

Haut-parleur à radiation directe et rayonnement optimisé

L'invention se rapporte à un haut-parleur à radiation directe du type comprenant classiquement une membrane mobile rattachée par sa périphérie extérieure à un châssis rigide formant cadre. Elle concerne plus particulièrement un perfectionnement permettant d'adapter la directivité d'un tel haut-parleur et plus particulièrement de reproduire la directivité d'un piston rectangulaire. Un intérêt de l'invention réside dans le fait que l'adaptation de directivité d'un tel haut-parleur permet de coupler plusieurs haut-parleurs en radiation directe, en supprimant les interférences sur une plage de fréquences étendue.

Un haut-parleur à radiation directe classique est constitué d'une membrane mobile relativement rigide, légère, conique ou à section exponentielle ou autre, au centre de laquelle est montée une bobine, mobile à l'intérieur d'un champ magnétique engendré par un aimant. La membrane mobile est rattachée par sa périphérie extérieure à un châssis rigide formant cadre qui constitue aussi le support de l'aimant. Le cadre coïncide avec ce qu'on désignera ci-dessous un plan d'émission acoustique au-delà duquel le son se propage dans le milieu extérieur. Un tel haut-parleur est l'un des composants les plus utilisés en sonorisation. Un signal électrique représentatif du son à reproduire est appliqué aux bornes de la bobine et celle-ci se déplace dans l'entrefer de l'aimant. Ce mouvement entraîne la membrane qui rayonne une énergie acoustique vers le milieu extérieur, au-delà dudit plan d'émission acoustique. Un tel haut-parleur présente les caractéristiques suivantes.

- Si le contour du cadre défini ci-dessus est circulaire, le rayonnement acoustique du haut-parleur est axisymétrique, c'est-à-dire identique dans tous les plans passant par l'axe du haut-parleur qui est aussi l'axe de la bobine mobile.
- La dispersion du haut-parleur diminue quand la fréquence augmente.

L'invention propose un accessoire destiné à être fixé à un tel haut-parleur pour modifier ses caractéristiques, en fonction de besoins spécifiques liés à la conception de l'enceinte acoustique, au moins pour une certaine plage de fréquences.

A cet effet, l'invention concerne un haut-parleur à radiation directe comprenant une membrane mobile rattachée par sa périphérie extérieure à un châssis rigide formant cadre, ce cadre définissant un plan d'émission acoustique, caractérisé en ce qu'il comporte en outre un obturateur de seulement une zone centrale dudit plan d'émission, à l'intérieur dudit cadre.

Ainsi, l'obturateur se trouve positionné devant une partie de la face avant du haut-parleur. Il se fixe mécaniquement sur le châssis ou sur une partie solidaire de celui-ci. Les moyens de fixation sont classiques, vis-écrou ou autres.

La forme de l'obturateur dépend généralement des résultats recherchés. Globalement cependant, l'obturateur est placé suivant un diamètre ou un axe de symétrie du cadre et recouvre typiquement entre le tiers et la moitié de la surface frontale dudit cadre, en laissant ouvertes deux parties égales dudit plan d'émission acoustique, symétriques par rapport à un axe de symétrie de l'obturateur.

Selon un mode de réalisation préféré, l'obturateur est défini dans une matière suffisamment rigide, éventuellement composite, pour ne pas être le siège de vibrations. Il peut par exemple être en matière plastique ou en bois. Il admet généralement au moins un plan de symétrie contenant un axe de la membrane qu'on appellera premier plan de symétrie et, de préférence, un second plan de symétrie contenant l'axe de la membrane et perpendiculaire audit premier plan de symétrie. Sa face arrière, c'est-à-dire celle qui est en regard de la membrane du haut-parleur, sera de préférence profilée. Ladite face arrière peut être par exemple globalement convexe de façon à s'engager dans l'espace défini entre la membrane et le plan d'émission acoustique.

L'invention sera mieux comprise et d'autres avantages de celle-ci apparaîtront plus clairement à la lumière de la description qui va suivre de plusieurs modes de réalisation d'un haut-parleur à radiation directe pourvu d'un adaptateur de directivité conforme à son principe, donnée uniquement à titre d'exemple et faite en référence aux dessins annexés dans lesquels :

- la figure 1 est une vue en perspective éclatée d'un haut-parleur à radiation directe et d'un adaptateur de directivité formant obturateur partiel ;

- la figure 2 est une vue analogue à la figure 1 montrant l'obturateur en place sur le cadre du châssis du haut-parleur ;

- la figure 3 est une vue de face dudit obturateur ;
- la figure 4 est une vue de profil en élévation du même obturateur ;
- la figure 5 est une vue en perspective de l'obturateur ;
- les figures 6 et 7 sont des schémas illustrant d'autres formes possibles d'obturateur ; et
- la figure 8 illustre le couplage de plusieurs haut-parleurs.

Sur les figures 1 à 5, on a représenté un haut-parleur à radiation directe 11 classique apte à recevoir un obturateur 25 constituant un adaptateur de directivité. Le haut-parleur comprend un châssis rigide 15 portant, à l'arrière, un aimant permanent 17 pourvu d'un entrefer cylindrique à l'intérieur duquel se déplace une bobine mobile solidaire d'une membrane mobile 19. La périphérie extérieure de la membrane est rattachée au châssis rigide et plus particulièrement à un cadre 21 de celui-ci, à contour intérieur circulaire. Le cadre comporte classiquement des trous 23 permettant la fixation du haut-parleur à une enceinte acoustique ou structure analogue.

On appelle ici "plan d'émission acoustique" le plan P contenant le contour du rattachement de la membrane au cadre du châssis. C'est à partir de ce plan que le son rayonne normalement dans l'air.

Selon une caractéristique importante de l'invention, le haut-parleur est en outre muni de l'obturateur 25 déjà mentionné, conformé pour obturer seulement une bande centrale dudit plan d'émission limité à l'intérieur dudit cadre. Par "seulement" on entend que cet obturateur est conformé pour laisser subsister deux larges ouvertures (figure 2) dans le plan d'émission acoustique P à l'intérieur du cadre de part et d'autre d'un premier plan de symétrie P1 contenant l'axe principal x'x de la membrane, qui est aussi l'axe de déplacement de sa bobine. La forme de ces deux ouvertures conjuguées et la forme de la face arrière de l'obturateur 25 permettent de redéfinir ou adapter les caractéristiques de dispersion de ce haut-parleur en radiation directe.

L'obturateur 25 est de structure rigide. Comme indiqué précédemment, il peut être en matière plastique, en bois, ou d'un autre matériau, éventuellement composite. Le matériau est choisi pour être le plus inerte possible, c'est-à-dire pour ne pas être le siège de vibrations parasites. Comme représenté, ledit premier plan de symétrie P1 contenant l'axe x'x est orienté suivant une direction

parallèle à la plus grande dimension de la bande centrale obturée. Préférentiellement, l'obturateur admet aussi un second plan de symétrie P2 contenant l'axe x'x de la membrane et perpendiculaire au premier plan de symétrie P1. Dans l'exemple, il comporte une face frontale 27 sensiblement plane. En revanche, sa face arrière 29, c'est-à-dire celle qui est tournée vers la membrane 19 du haut-parleur, est de préférence profilée. Par exemple, comme représenté, ladite face arrière est globalement convexe et s'engage dans l'espace défini entre la membrane 19 et le plan d'émission acoustique P. Plus précisément, elle est définie par l'intersection d'une surface convexe bombée 31 et de deux échancrures latérales 33 courbes et concaves s'étendant de part et d'autre du premier plan de symétrie P1. Les deux échancrures latérales courbes sont symétriques par rapport audit premier plan de symétrie. Chacune d'elles est symétrique par rapport audit second plan de symétrie.

Selon une autre caractéristique remarquable, bien que facultative, la surface bombée 31 a sensiblement la même forme que la partie de la membrane en regard de laquelle elle se trouve. Autrement dit, la surface bombée est sensiblement en tous points, à la même distance de la membrane.

La face frontale 27 est globalement rectangulaire bien que les deux échancrures 33 définissent un rétrécissement dans sa partie médiane. L'obturateur comporte quatre trous de fixation 35 espacés pour venir en correspondance avec quatre trous 23 du cadre du châssis.

En fonctionnement normal, le haut-parleur est disposé comme représenté à la figure 2, c'est-à-dire de façon que la bande centrale recouverte par l'obturateur 25 soit sensiblement verticale. Dans cette configuration, l'obturateur élargit la dispersion dans le plan vertical et la réduit dans le plan horizontal. On a trouvé que la forme décrite ci-dessus permet d'adapter de façon favorable les caractéristiques de dispersion du haut-parleur sans affecter de façon sensible ses autres performances intrinsèques, notamment en ce qui concerne le rendement, la puissance admissible et le taux de distorsion.

Tel que représenté, cet obturateur, du fait qu'il réduit la couverture dans le plan horizontal, permet de coupler horizontalement plusieurs haut-parleurs à radiation directe, en supprimant les interférences sur une plage de fréquences étendue.

La figure 8 montre comment on peut inclure un tel haut-parleur 11 muni de son obturateur 25 dans une enceinte acoustique spécifique 40. Chaque enceinte a une section horizontale trapézoïdale. Les enceintes acoustiques sont accolées par leurs faces latérales. Ce type de montage permet de coupler les haut-parleurs 11 sans provoquer d'interférences entre eux.

Les figures 6 et 7 illustrent d'autres variantes possibles. Dans le cas de la figure 6, l'obturateur 25a est réduit à une forme très simple, il s'agit d'une plaque rectangulaire venant recouvrir seulement une bande centrale du plan d'émission. Dans le mode de réalisation de la figure 7, la face frontale se compose de l'association de deux portions rectangulaires 37 situées de part et d'autre d'une portion de disque 39. Pour chacun de ces deux modes de réalisation, la face arrière 29 peut être plane ou, de préférence, profilée de façon comparable à ce qui a été décrit en référence aux figures 3 à 5.

REVENDICATIONS

1. Haut-parleur à radiation directe comprenant une membrane mobile (19) rattachée par sa périphérie extérieure à un châssis rigide (15) formant cadre, ce cadre définissant un plan d'émission acoustique (P), caractérisé en ce qu'il comporte en outre un obturateur (25) de seulement une zone centrale dudit plan d'émission à l'intérieur dudit cadre.

2. Haut-parleur selon la revendication 1, caractérisé en ce que ledit obturateur (25), de structure rigide, admet un premier plan de symétrie (P1) contenant un axe (x'x) de ladite membrane.

3. Haut-parleur selon la revendication 2, caractérisé en ce que ledit obturateur admet un second plan de symétrie (P2) contenant ledit axe de la membrane et perpendiculaire audit premier plan de symétrie.

4. Haut-parleur selon la revendication 3, caractérisé en ce que ledit obturateur comporte une face frontale (27) sensiblement plane.

5. Haut-parleur selon la revendication 3 ou 4, caractérisé en ce que ledit obturateur a une face arrière (29) profilée.

6. Haut-parleur selon la revendication 4 ou 5, caractérisé en ce que ladite face frontale est globalement rectangulaire.

7. Haut-parleur selon la revendication 5 ou 6, caractérisé en ce que ladite face arrière (29) est globalement convexe et s'engage dans l'espace défini entre ladite membrane (19) et ledit plan d'émission acoustique (P).

8. Haut-parleur selon la revendication 7, caractérisé en ce que ladite face arrière est définie par l'intersection d'une surface bombée (31) et de deux échancrures latérales courbes (33) s'étendant de part et d'autre de l'un desdits plans de symétrie.

9. Haut-parleur selon la revendication 8, caractérisé en ce que ladite surface bombée (31) a sensiblement la même forme que la partie de la membrane en regard de laquelle elle se trouve.

Fig.1

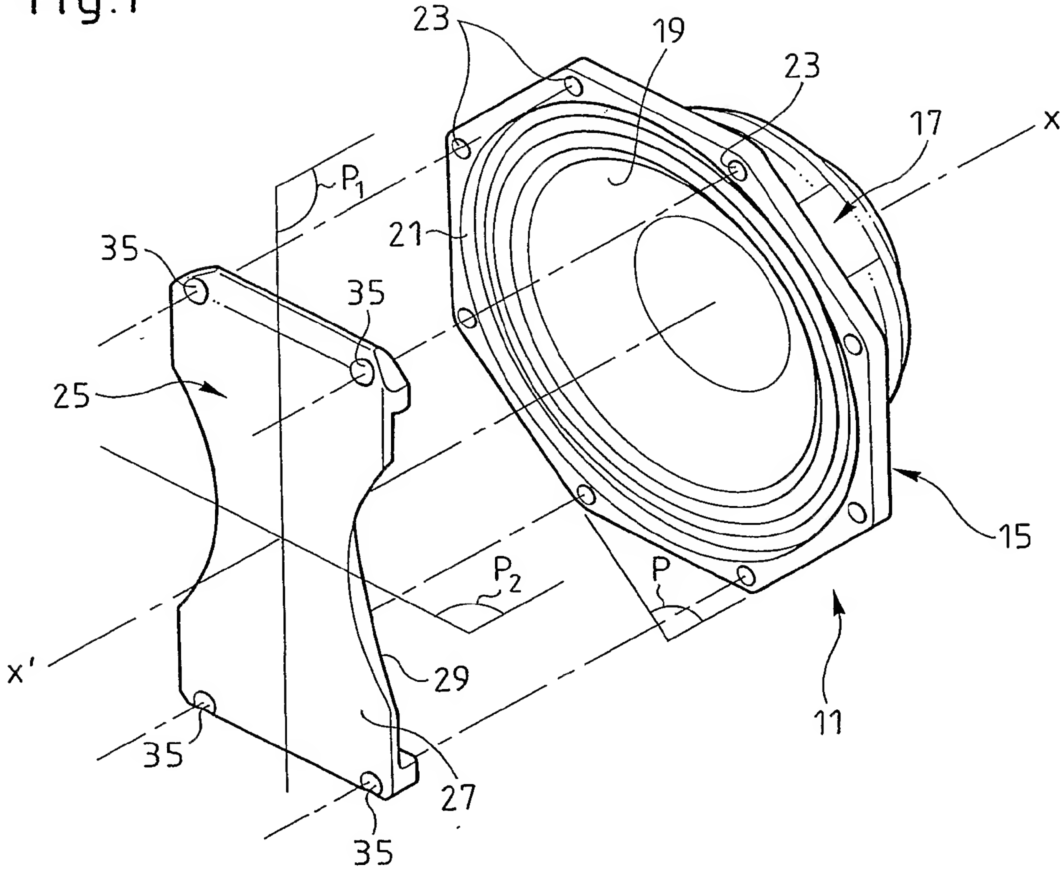

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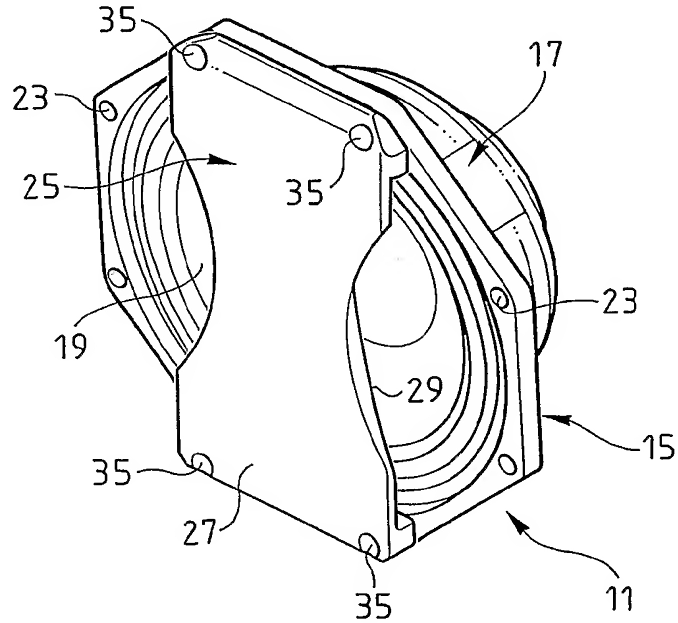
Fig.2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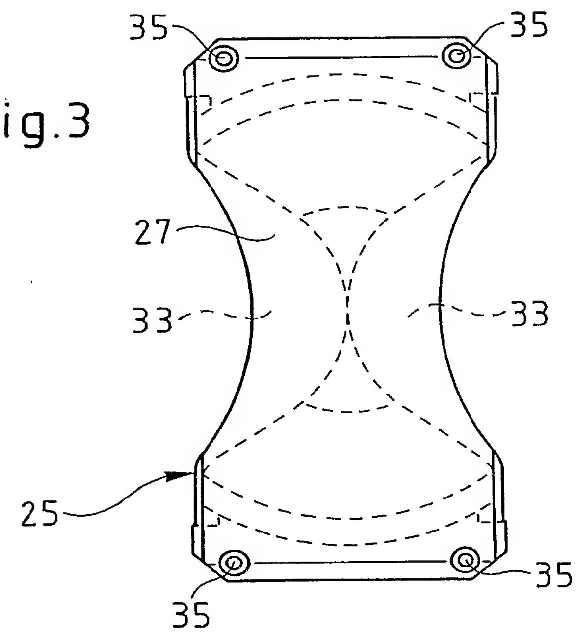

Fig.3

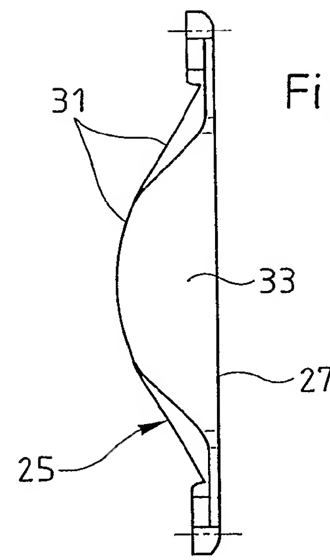

Fig.4

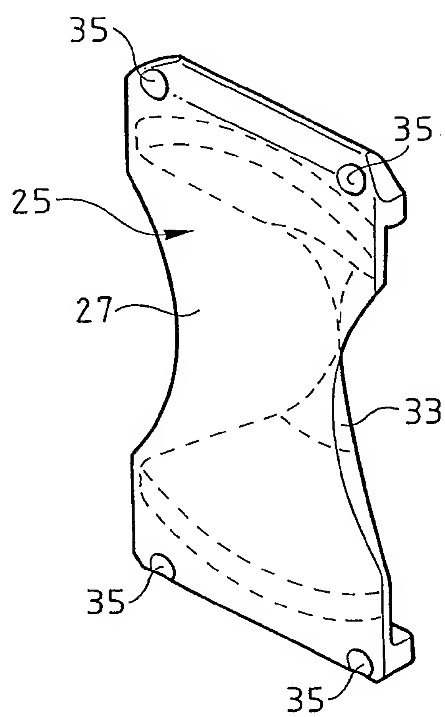

Fig.5

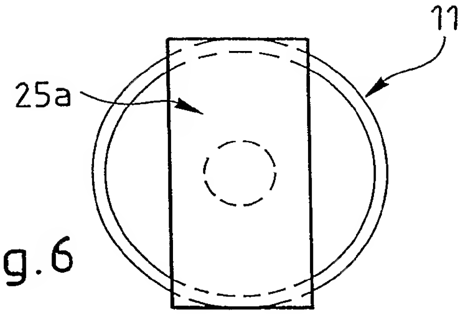

Fig.6

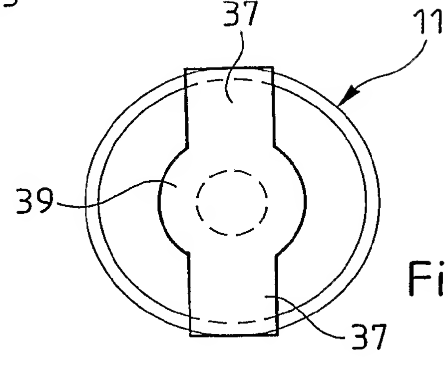

Fig.7

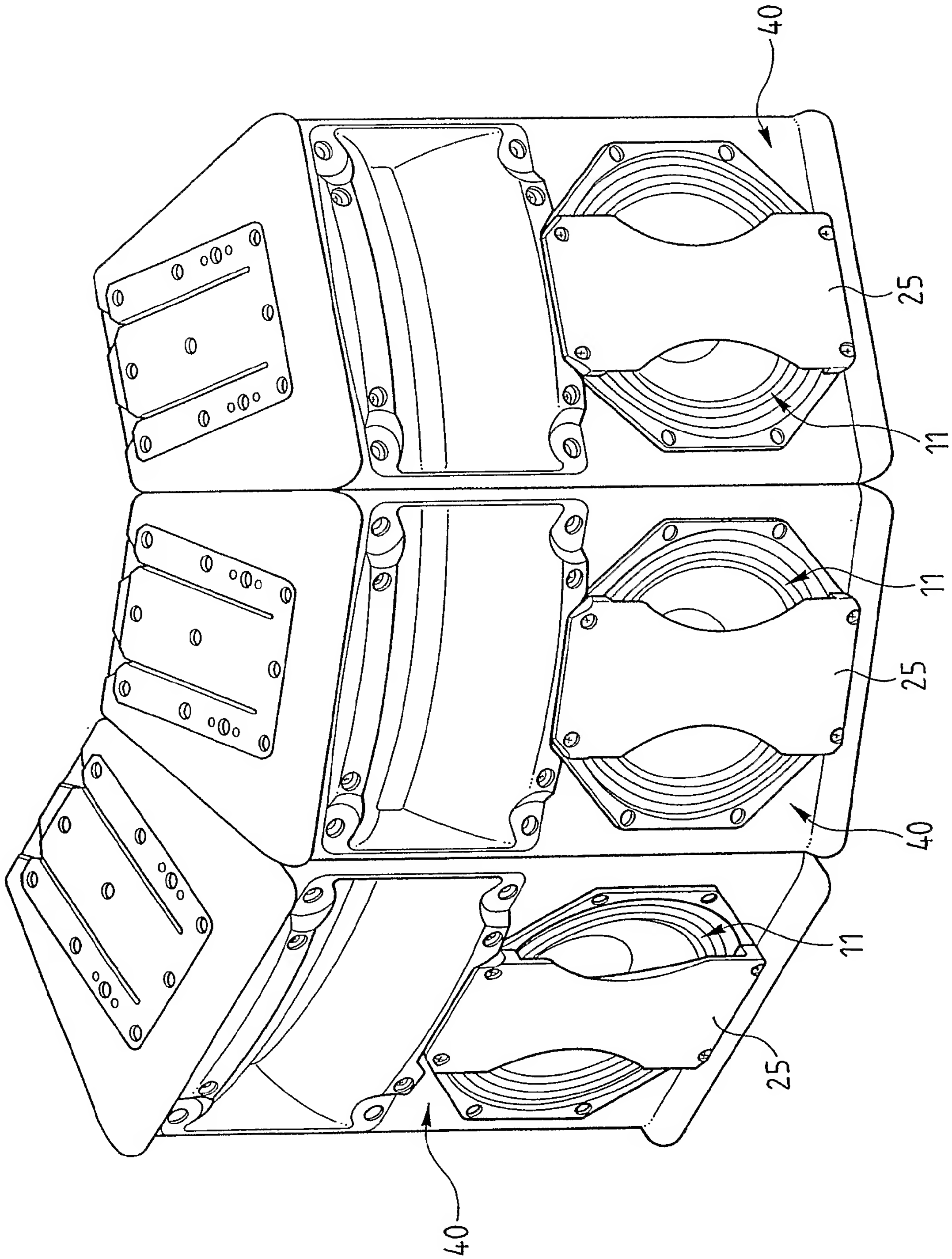

Fig.8

INTERNATIONAL SEARCH REPORT

International Application No
PCT/FR 02/02615

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04R1/34

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04R

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

WPI Data, PAJ

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	DE 42 11 114 A (EHMANN ELEKTROTECH) 7 October 1993 (1993-10-07) column 3, line 3-26	1-4,6
A	figure 2	5,7-9
X	GB 2 027 320 A (DITCHBURN ORGANISATION LTD) 13 February 1980 (1980-02-13) page 1, line 72-94	1-4,6
A	figure 2	5,7-9
A	PATENT ABSTRACTS OF JAPAN vol. 013, no. 548 (E-856), 7 December 1989 (1989-12-07) -& JP 01 226298 A (MATSUSHITA ELECTRIC IND CO LTD), 8 September 1989 (1989-09-08) abstract	1-9

☐ Further documents are listed in the continuation of box C.

☒ Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
22 November 2002	29/11/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Zanti, P

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/FR 02/02615

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
DE 4211114	A	07-10-1993	DE	4211114 A1	07-10-1993
GB 2027320	A	13-02-1980	NONE		
JP 01226298	A	08-09-1989	NONE		

RAPPORT DE RECHERCHE INTERNATIONALE

De...de Internationale No
PCT/FR 02/02615

A. CLASSEMENT DE L'OBJET DE LA DEMANDE
CIB 7 H04R1/34

Selon la classification internationale des brevets (CIB) ou à la fois selon la classification nationale et la CIB

B. DOMAINES SUR LESQUELS LA RECHERCHE A PORTE

Documentation minimale consultée (système de classification suivi des symboles de classement)
CIB 7 H04R

Documentation consultée autre que la documentation minimale dans la mesure où ces documents relèvent des domaines sur lesquels a porté la recherche

Base de données électronique consultée au cours de la recherche internationale (nom de la base de données, et si réalisable, termes de recherche utilisés)

WPI Data, PAJ

C. DOCUMENTS CONSIDERES COMME PERTINENTS

Catégorie °	Identification des documents cités, avec, le cas échéant, l'indication des passages pertinents	no. des revendications visées
X	DE 42 11 114 A (EHMANN ELEKTROTECH) 7 octobre 1993 (1993-10-07) colonne 3, ligne 3-26	1-4,6
A	figure 2	5,7-9
X	GB 2 027 320 A (DITCHBURN ORGANISATION LTD) 13 février 1980 (1980-02-13) page 1, ligne 72-94	1-4,6
A	figure 2	5,7-9
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(71) Applicant: QUALCOMM INCORPORATED [US/US]; 5775 Morehouse Drive, San Diego, CA 92121 (US).

(72) Inventor: KADOUS, Tamer; 5385 Toscana Way, #316, San Diego, CA 92122 (US).

(74) Agent: WADSWORTH, Philip R.; QUALCOMM Incorporated, 5775 Morehouse Drive, San Diego, CA 92121 (US).

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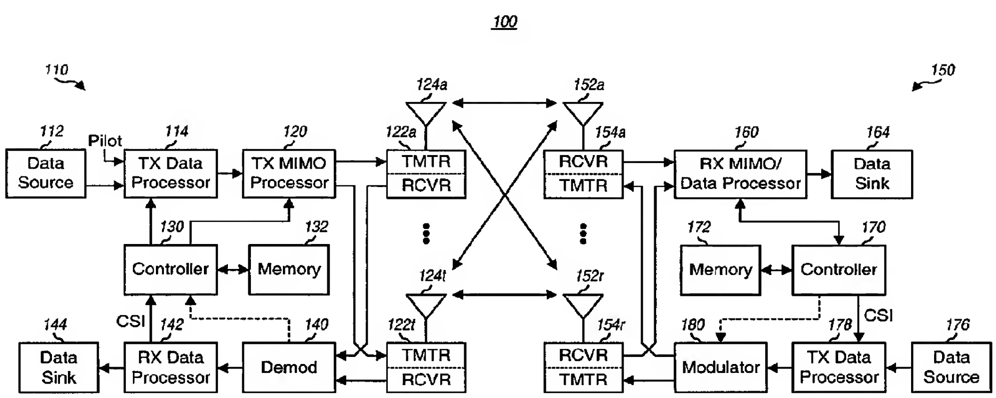
(54) **Title:** DATA TRANSMISSION WITH NON-UNIFORM DISTRIBUTION OF DATA RATES FOR A MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) SYSTEM



(57) **Abstract:** Techniques to determine data rates for a number of data streams transmitted via a number of transmission channels (or transmit antennas) in a multi-channel (e.g., MIMO) communication system. In one method, the "required" SNR for each data rate to be used is initially determined, with at least two data rates being unequal. The "effective" SNR for each data stream is also determined based on the received SNR and successive interference cancellation processing at the receiver to recover the data streams. The required SNR for each data stream is then compared against its effective SNR. The data rates are deemed to be supported if the required SNR for each data stream is less than or equal to its effective SNR. A number of sets of data rates may be evaluated, and the rate set associated with the minimum received SNR may be selected for use for the data streams.

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DATA TRANSMISSION WITH NON-UNIFORM DISTRIBUTION OF DATA RATES FOR A MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) SYSTEM

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for determining a non-uniform distribution of data rates to be used for multiple data streams to be transmitted via multiple transmission channels of a multi-channel communication system, e.g., a multiple-input multiple-output (MIMO) system.

Background

[1002] In a wireless communication system, an RF modulated signal from a transmitter may reach a receiver via a number of propagation paths. The characteristics of the propagation paths typically vary over time due to a number of factors such as fading and multipath. To provide diversity against deleterious path effects and improve performance, multiple transmit and receive antennas may be used. If the propagation paths between the transmit and receive antennas are linearly independent (i.e., a transmission on one path is not formed as a linear combination of the transmissions on the other paths), which is generally true to at least an extent, then the likelihood of correctly receiving a data transmission increases as the number of antennas increases. Generally, diversity increases and performance improves as the number of transmit and receive antennas increases.

[1003] A multiple-input multiple-output (MIMO) communication system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels may also be referred to as a spatial subchannel (or a transmission channel) of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission

capacity) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1004] For a full-rank MIMO channel, where $N_S = N_T \le N_R$, an independent data stream may be transmitted from each of the N_T transmit antennas. The transmitted data streams may experience different channel conditions (e.g., different fading and multipath effects) and may achieve different signal-to-noise-and-interference ratios (SNRs) for a given amount of transmit power. Moreover, if successive interference cancellation processing is used at the receiver to recover the transmitted data streams (described below), then different SNRs may be achieved for the data streams depending on the specific order in which the data streams are recovered. Consequently, different data rates may be supported by different data streams, depending on their achieved SNRs. Since the channel conditions typically vary with time, the data rate supported by each data stream also varies with time.

[1005] If the characteristics of the MIMO channel (e.g., the achieved SNRs for the data streams) are known at the transmitter, then the transmitter may be able to determine a particular data rate and coding and modulation scheme for each data stream such that an acceptable level of performance (e.g., one percent packet error rate) may be achieved for the data stream. However, for some MIMO systems, this information is not available at the transmitter. Instead, what may be available is very limited amount of information regarding, for example, the operating SNR for the MIMO channel, which may be defined as the expected SNR for all data streams at the receiver. In this case, the transmitter would need to determine the proper data rate and coding and modulation scheme for each data stream based on this limited information.

[1006] There is therefore a need in the art for techniques to determine a set of data rates for multiple data streams to achieve high performance when limited information is available at the transmitter for the MIMO channel.

SUMMARY

[1007] Techniques are provided herein to provide improved performance for a MIMO system when channel state information indicative of the current channel conditions is not available at the transmitter. In an aspect, a non-uniform distribution of data rates is used for the transmitted data streams. The data rates may be selected to achieve (1) a specified overall spectral efficiency with a lower minimum "received"

SNR (described below) or (2) a higher overall spectral efficiency for a specified received SNR. A specific scheme for achieving each of the above objectives is provided herein.

[1008] In a specific embodiment that may be used to achieve the first objective noted above, a method is provided for determining data rates to be used for a number of data streams to be transmitted via a number of transmission channels in a multi-channel communication system (e.g., one data stream may be transmitted over each transmit antenna in a MIMO system). In accordance with the method, the required SNR for each of a number of data rates to be used for the data streams is initially determined. At least two of the data rates are unequal. The "effective" SNR (described below) for each data stream is also determined based on the received SNR and successive interference cancellation processing at the receiver (also described below) to recover the data streams. The required SNR for each data stream is then compared against the effective SNR for the data stream. The data rates are deemed to be supported if the required SNR for each data stream is less than or equal to the effective SNR for the data stream. A number of sets of data rates may be evaluated, and the rate set associated with the minimum received SNR may be selected for use for the data streams.

[1009] In a specific embodiment that may be used to achieve the second objective noted above, a method is provided for determining data rates for a number of data streams to be transmitted via a number of transmission channels (e.g., transmit antennas) in a multi-channel (e.g., MIMO) communication system. In accordance with the method, the received SNR is initially determined. This received SNR may be specified for the system or may be estimated based on measurements at the receiver and periodically provided to the transmitter. The effective SNR for each data stream is also determined based on the received SNR and successive interference cancellation processing at the receiver. The data rate for each data stream is then determined based on the effective SNR for the data stream, with at least two of the data rates being unequal.

[1010] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, processors, transmitter units, receiver units, base stations, terminals, systems, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1011] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1012] FIG. 1 is a block diagram of an embodiment of a transmitter system and a receiver system in a MIMO system;

[1013] FIG. 2 is a flow diagram illustrating a successive interference cancellation receiver processing technique to process N_R received symbol streams to recover N_T transmitted symbol streams;

[1014] FIG. 3 is a flow diagram of an embodiment of a process for determining the minimum received SNR needed to support a given set of data rates;

[1015] FIG. 4 shows plots of packet error rate (PER) versus SNR for a {1, 4} MIMO system for spectral efficiencies of 1, 4/3, 5/3, and 2 bps/Hz;

[1016] FIG. 5 is a block diagram of an embodiment of a transmitter unit; and

[1017] FIG. 6 is a block diagram of an embodiment of a receiver unit capable of implementing the successive interference cancellation receiver processing technique.

DETAILED DESCRIPTION

[1018] The techniques described herein for determining a set of data rates for multiple data streams based on limited channel state information may be implemented in various multi-channel communication systems. Such multi-channel communication systems include multiple-input multiple-output (MIMO) communication systems, orthogonal frequency division multiplexing (OFDM) communication systems, MIMO systems that employ OFDM (i.e., MIMO-OFDM systems), and so on. For clarity, various aspects and embodiments are described specifically for a MIMO system.

[1019] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels may also be referred to as a spatial subchannel (or transmission channel) of the MIMO channel. The number of spatial subchannels is determined by the number of eigenmodes for the MIMO channel,

which in turn is dependent on a channel response matrix, $\underline{\mathbf{H}}$, that describes the response between the N_T transmit and N_R receive antennas. The elements of the channel response matrix, $\underline{\mathbf{H}}$, are composed of independent Gaussian random variables $\{h_{i,j}\}$, for $i=1, 2, \ldots N_R$ and $j=1, 2, \ldots N_T$, where $h_{i,j}$ is the coupling (i.e., the complex gain) between the j-th transmit antenna and the i-th receive antenna. For simplicity, the channel response matrix, $\underline{\mathbf{H}}$, is assumed to be full-rank (i.e., $N_S = N_T \leq N_R$), and one independent data stream may be transmitted from each of the N_T transmit antennas.

[1020] FIG. 1 is a block diagram of an embodiment of a transmitter system 110 and a receiver system 150 in a MIMO system 100.

[1021] At transmitter system 110, traffic data for a number of data streams is provided from a data source 112 to a transmit (TX) data processor 114. In an embodiment, each data stream is transmitted over a respective transmit antenna. TX data processor 114 formats, codes, and interleaves the traffic data for each data stream based on a particular coding scheme selected for that data stream to provide coded data.

[1022] The coded data for each data stream may be multiplexed with pilot data using, for example, time division multiplexing (TDM) or code division multiplexing (CDM). The pilot data is typically a known data pattern that is processed in a known manner (if at all), and may be used at the receiver system to estimate the channel response. The multiplexed pilot and coded data for each data stream is then modulated (i.e., symbol mapped) based on a particular modulation scheme (e.g., BPSK, QSPK, M-PSK, or M-QAM) selected for that data stream to provide modulation symbols. The data rate, coding, and modulation for each data stream may be determined by controls provided by a controller 130.

[1023] The modulation symbols for all data streams are then provided to a TX MIMO processor 120, which may further process the modulation symbols (e.g., for OFDM). TX MIMO processor 120 then provides N_T modulation symbol streams to N_T transmitters (TMTR) 122a through 122t. Each transmitter 122 receives and processes a respective symbol stream to provide one or more analog signals, and further conditions (e.g., amplifies, filters, and upconverts) the analog signals to provide a modulated signal suitable for transmission over the MIMO channel. N_T modulated signals from transmitters 122a through 122t are then transmitted from N_T antennas 124a through 124t, respectively.

[1024] At receiver system 150, the transmitted modulated signals are received by N_R antennas 152a through 152r, and the received signal from each antenna 152 is provided to a respective receiver (RCVR) 154. Each receiver 154 conditions (e.g., filters, amplifies, and downconverts) a respective received signal, digitizes the conditioned signal to provide samples, and further processes the samples to provide a corresponding "received" symbol stream.

[1025] An RX MIMO/data processor 160 then receives and processes the N_R received symbol streams from N_R receivers 154 based on a particular receiver processing technique to provide N_T "detected" symbol streams. The processing by RX MIMO/data processor 160 is described in further detail below. Each detected symbol stream includes symbols that are estimates of the modulation symbols transmitted for the corresponding data stream. RX MIMO/data processor 160 then demodulates, deinterleaves, and decodes each detected symbol stream to recover the traffic data for the data stream. The processing by RX MIMO/data processor 160 is complementary to that performed by TX MIMO processor 120 and TX data processor 114 at transmitter system 110.

[1026] RX MIMO processor 160 may derive an estimate of the channel response between the N_T transmit and N_R receive antennas, e.g., based on the pilot multiplexed with the traffic data. The channel response estimate may be used to perform space or space/time processing at the receiver. RX MIMO processor 160 may further estimate the signal-to-noise-and-interference ratios (SNRs) of the detected symbol streams, and possibly other channel characteristics, and provides these quantities to a controller 170. RX MIMO/data processor 160 or controller 170 may further derive an estimate of the "operating" SNR for the system, which is indicative of the conditions of the communication link. Controller 170 then provides channel state information (CSI), which may comprise various types of information regarding the communication link and/or the received data stream. For example, the CSI may comprise only the operating SNR. The CSI is then processed by a TX data processor 178, modulated by a modulator 180, conditioned by transmitters 154a through 154r, and transmitted back to transmitter system 110.

[1027] At transmitter system 110, the modulated signals from receiver system 150 are received by antennas 124, conditioned by receivers 122, demodulated by a demodulator 140, and processed by a RX data processor 142 to recover the CSI reported

by the receiver system. The reported CSI is then provided to controller 130 and used to (1) determine the data rates and coding and modulation schemes to be used for the data streams and (2) generate various controls for TX data processor 114 and TX MIMO processor 120.

[1028] Controllers 130 and 170 direct the operation at the transmitter and receiver systems, respectively. Memories 132 and 172 provide storage for program codes and data used by controllers 130 and 170, respectively.

[1029] The model for the MIMO system may be expressed as:

$$\underline{\mathbf{y}} = \underline{\mathbf{H}}\underline{\mathbf{x}} + \underline{\mathbf{n}} \ , \qquad \text{Eq (1)}$$

where $\underline{\mathbf{y}}$ is the received vector, i.e., $\underline{\mathbf{y}} = [y_1 \ y_2 \ldots \ y_{N_R}]^T$, where $\{y_i\}$ is the entry received on the i-th received antenna and $i \in \{1, \ldots, N_R\}$;

$\underline{\mathbf{x}}$ is the transmitted vector, i.e., $\underline{\mathbf{x}} = [x_1 \ x_2 \ldots \ x_{N_T}]^T$, where $\{x_j\}$ is the entry transmitted from the j-th transmit antenna and $j \in \{1, \ldots, N_T\}$;

$\underline{\mathbf{H}}$ is the channel response matrix for the MIMO channel;

$\underline{\mathbf{n}}$ is the additive white Gaussian noise (AWGN) with a mean vector of $\underline{\mathbf{0}}$ and a covariance matrix of $\underline{\mathbf{\Lambda}}_n = \sigma^2 \underline{\mathbf{I}}$, where $\underline{\mathbf{0}}$ is a vector of zeros, $\underline{\mathbf{I}}$ is the identity matrix with ones along the diagonal and zeros everywhere else, and σ^2 is the variance of the noise; and

$[.]^T$ denotes the transpose of $[.]$.

[1030] Due to scattering in the propagation environment, the N_T symbol streams transmitted from the N_T transmit antennas interfere with each other at the receiver. In particular, a given symbol stream transmitted from one transmit antenna may be received by all N_R receive antennas at different amplitudes and phases. Each received signal may then include a component of each of the N_T transmitted symbol streams. The N_R received signals would collectively include all N_T transmitted symbols streams. However, these N_T symbol streams are dispersed among the N_R received signals.

[1031] At the receiver, various processing techniques may be used to process the N_R received signals to detect the N_T transmitted symbol streams. These receiver processing techniques may be grouped into two primary categories:

- spatial and space-time receiver processing techniques (which are also referred to as equalization techniques), and
- "successive nulling/equalization and interference cancellation" receiver processing technique (which is also referred to as "successive interference cancellation" or "successive cancellation" receiver processing technique).

[1032] In general, the spatial and space-time receiver processing techniques attempt to separate out the transmitted symbol streams at the receiver. Each transmitted symbol stream may be "detected" by (1) combining the various components of the transmitted symbol stream included in the N_R received signals based on an estimate of the channel response and (2) removing (or canceling) the interference due to the other symbol streams. These receiver processing techniques attempt to either (1) decorrelate the individual transmitted symbol streams such that there is no interference from the other symbol streams or (2) maximize the SNR of each detected symbol stream in the presence of noise and interference from the other symbol streams. Each detected symbol stream is then further processed (e.g., demodulated, deinterleaved, and decoded) to recover the traffic data for the symbol stream.

[1033] The successive cancellation receiver processing technique attempts to recover the transmitted symbol streams, one at a time, using spatial or space-time receiver processing, and to cancel the interference due to each "recovered" symbol stream such that later recovered symbol streams experience less interference and may be able to achieve higher SNRs. The successive cancellation receiver processing technique may be used if the interference due to each recovered symbol stream can be accurately estimated and canceled, which requires error-free or low-error recovery of the symbol stream. The successive cancellation receiver processing technique (which is described in further detail below) generally outperforms the spatial/space-time receiver processing techniques.

[1034] For the successive cancellation receiver processing technique, the N_R received symbol streams are processed by N_T stages to successively recover one transmitted symbol stream at each stage. As each transmitted symbol stream is recovered, the interference it causes to the remaining not yet recovered symbol streams is estimated and canceled from the received symbol streams, and the "modified" symbol streams are further processed by the next stage to recover the next transmitted symbol stream. If the transmitted symbol streams can be recovered without error (or with

minimal errors) and if the channel response estimate is reasonably accurate, then cancellation of the interference due to the recovered symbol stream is effective, and the SNR of each subsequently recovered symbol stream is improved. In this way, higher performance may be achieved for all transmitted symbol streams (possibly except for the first transmitted symbol stream to be recovered).

[1035] The following terminology is used herein:

- "transmitted" symbol streams - the symbol streams transmitted from the transmit antennas;
- "received" symbol streams - the inputs to a spatial or space-time processor in the first stage of a successive interference cancellation (SIC) receiver (see FIG. 6);
- "modified" symbol streams - the inputs to the spatial or space-time processor in each subsequent stage of the SIC receiver;
- "detected" symbol streams - the outputs from the spatial processor (up to $N_T - k + 1$ symbol streams may be detected at stage *k*); and
- "recovered" symbol stream - a symbol stream that has been decoded at the receiver (only one detected symbol stream is recovered at each stage).

[1036] FIG. 2 is a flow diagram illustrating the successive cancellation receiver processing technique to process N_R received symbol streams to recover N_T transmitted symbol streams. For simplicity, the following description for FIG. 2 assumes that (1) the number of spatial subchannels is equal to the number of transmit antennas (i.e., $N_S = N_T \le N_R$) and (2) one independent data stream is transmitted from each transmit antenna.

[1037] For the first stage ($k = 1$), the receiver initially performs spatial or space-time processing on the N_R received symbol streams to attempt to separate out the N_T transmitted symbol streams (step 212). For the first stage, the spatial or space-time processing can provide N_T detected symbol streams that are estimates of the N_T (not yet recovered) transmitted symbol streams. One of the detected symbol streams is then selected (e.g., based on a particular selection scheme) and further processed. If the identity of the transmitted symbol stream to be recovered in the stage is known *a priori*, then the space or space-time processing may be performed to provide only one detected symbol stream for this transmitted symbol stream. In either case, the selected detected symbol stream is further processed (e.g., demodulated, deinterleaved, and decoded) to

obtain a decoded data stream, which is an estimate of the data stream for the transmitted symbol stream being recovered in this stage (step 214).

[1038] A determination is then made whether or not all transmitted symbol streams have been recovered (step 216). If the answer is yes, then the receiver processing terminates. Otherwise, the interference due to the just-recovered symbol stream on each of the N_R received symbol streams is estimated (step 218). The interference may be estimated by first re-encoding the decoded data stream, interleaving the re-encoded data, and symbol mapping the interleaved data (using the same coding, interleaving, and modulation schemes used at the transmitter unit for this data stream) to obtain a "remodulated" symbol stream, which is an estimate of the transmitted symbol stream just recovered. The remodulated symbol stream is then convolved by each of N_R elements in a channel response vector $\underline{\mathbf{h}}_j$ to derive N_R interference components due to the just-recovered symbol stream. The vector $\underline{\mathbf{h}}_j$ is a column of the $(N_R \times N_T)$ channel response matrix, $\underline{\mathbf{H}}$, corresponding to the j-th transmit antenna used for the just-recovered symbol stream. The vector $\underline{\mathbf{h}}_j$ includes N_R elements that define the channel response between the j-th transmit antenna and the N_R receive antennas

[1039] The N_R interference components are then subtracted from the N_R received symbol streams to derive N_R modified symbol streams (step 220). These modified symbol streams represent the streams that would have been received if the just-recovered symbol stream had not been transmitted (i.e., assuming that the interference cancellation was effectively performed).

[1040] The processing performed in steps 212 and 214 is then repeated on the N_R modified symbol streams (instead of the N_R received symbol streams) to recover another transmitted symbol stream. Steps 212 and 214 are thus repeated for each transmitted symbol stream to be recovered, and steps 218 and 220 are performed if there is another transmitted symbol stream to be recovered.

[1041] For the first stage, the input symbol streams are the N_R received symbol streams from the N_R received antennas. And for each subsequent stage, the input symbol streams are the N_R modified symbol streams from the preceding stage. The processing for each stage proceeds in similar manner. At each stage subsequent to the first stage, the symbol streams recovered in the prior stages are assumed to be cancelled,

so the dimensionality of the channel response matrix $\underline{\mathbf{H}}$ is successively reduced by one column for each subsequent stage.

[1042] The successive cancellation receiver processing thus includes a number of stages, one stage for each transmitted symbol stream to be recovered. Each stage recovers one of the transmitted symbol streams and (except for the last stage) cancels the interference due to this recovered symbol stream to derive the modified symbol streams for the next stage. Each subsequently recovered symbol stream thus experiences less interference and is able to achieve a higher SNR than without the interference cancellation. The SNRs of the recovered symbol streams are dependent on the particular order in which the symbol streams are recovered.

[1043] For the successive cancellation receiver processing, the input symbol streams for the k-th stage (assuming that the interference from the symbol streams recovered in the prior k-1 stages have been effectively canceled) may be expressed as:

$$\underline{\mathbf{y}}_k = \underline{\mathbf{H}}_k \underline{\mathbf{x}}_k + \underline{\mathbf{n}} \ , \qquad \text{Eq (2)}$$

where $\underline{\mathbf{y}}_k$ is the $N_R \times 1$ input vector for the k-th stage, i.e., $\underline{\mathbf{y}}_k = [y_1^{\ k} \ \ y_2^{\ k} \ \ldots \ y_{N_R}^k]^T$, where $y_i^{\ k}$ is the entry for the i-th received antenna in the k-th stage;

$\underline{\mathbf{x}}_k$ is the $(N_T - k + 1) \times 1$ transmitted vector for the k-th stage, i.e., $\underline{\mathbf{x}}_k = [x_k \ \ x_{k+1} \ldots x_{N_T}]^T$, where x_j is the entry transmitted from the j-th transmit antenna;

$\underline{\mathbf{H}}_k$ is the $N_R \times (N_T - k + 1)$ channel response matrix for the MIMO channel, with k-1 columns for the previously recovered symbol streams removed, i.e., $\underline{\mathbf{H}}_k = [\underline{\mathbf{h}}_k \ \ \underline{\mathbf{h}}_{k+1} \ \ldots \ \underline{\mathbf{h}}_{N_T}]$; and

$\underline{\mathbf{n}}$ is the additive white Gaussian noise

For simplicity, equation (2) assumes that the transmitted symbol streams are recovered in the order of the transmit antennas (i.e., the symbol stream transmitted from transmit antenna 1 is recovered first, the symbol stream transmitted from transmit antenna 2 is recovered second, and so on, and the symbol stream transmitted from transmit antenna N_T is recovered last). Equation (2) may be rewritten as:

$$\underline{\mathbf{y}}_k = \sum_{j=k}^{N_T} \underline{\mathbf{h}}_j \underline{\mathbf{x}}_j + \underline{\mathbf{n}} \quad . \tag{Eq (3)}$$

[1044] The transmitted symbol stream to be recovered in stage k may be viewed as being projected at a particular angle from an interference sub-space (or plane) $\underline{\mathbf{S}}^I$. The transmitted symbol stream is dependent on (and defined by) the channel response vector $\underline{\mathbf{h}}_k$. An interference-free component of the transmitted symbol stream may be obtained by projecting the channel response vector, $\underline{\mathbf{h}}_k$, on an interference-free sub-space, which is orthogonal to the interference sub-space. This projection may be achieved by multiplying $\underline{\mathbf{h}}_k$ with a filter having a response of $\underline{\mathbf{w}}$. The filter that attains the maximum energy after the projection is the one that lies in a sub-space constructed by $\underline{\mathbf{h}}_k$ and the interference sub-space $\underline{\mathbf{S}}^I$, where $\underline{\mathbf{S}}^I = \text{span}\ (\underline{\mathbf{i}}_1\ \underline{\mathbf{i}}_2\ \ldots\ \underline{\mathbf{i}}_{N_T-k})$, $\underline{\mathbf{i}}_m^H \underline{\mathbf{i}}_n = \delta_{m,n}$, and $\{\underline{\mathbf{i}}_n\}$, for $n = 1,\ 2,\ \ldots\ N_T - k$, are orthonormal basis spanning the interference sub-space $\underline{\mathbf{S}}^I$. The average energy after the projection is given by:

$$\begin{aligned} E[|\underline{\mathbf{w}}^H \underline{\mathbf{h}}_k|^2] &= E[|\underline{\mathbf{h}}_k^H \underline{\mathbf{h}}_k|^2] - E[|\underline{\mathbf{S}}^{I^H} \underline{\mathbf{h}}_k|^2] \\ &= \frac{N_R}{N_T} - \sum_{j=1}^{N_T-k} \underline{\mathbf{i}}_j^H E[|\underline{\mathbf{h}}_k \underline{\mathbf{h}}_k^H|^2] \underline{\mathbf{i}}_j \\ &= \frac{N_R - N_T + k}{N_T} \quad , \end{aligned} \tag{Eq (4)}$$

where $\underline{\mathbf{w}}^H \underline{\mathbf{h}}_k$ represents the projection of $\underline{\mathbf{h}}_k$ on the interference-free sub-space (i.e., the desired component), and

$\underline{\mathbf{S}}^{I^H} \underline{\mathbf{h}}_k$ represents the projection of $\underline{\mathbf{h}}_k$ on the interference sub-space (i.e., the interference component).

Equation (4) assumes equal transmit powers being used for the transmit antennas.

[1045] The effective SNR for the symbol stream recovered in the k-th stage, $\text{SNR}_{\text{eff}}(k)$, may be expressed as:

$$\text{SNR}_{\text{eff}}(k) = \frac{P_{tot}(N_R - N_T + k)}{\sigma^2 N_T} \quad , \tag{Eq (5)}$$

where P_{tot} is the total transmit power available for data transmission, which is uniformly distributed across the N_T transmit antennas such that P_{tot} / N_T is used for each transmit antenna, and

σ^2 is the noise variance.

[1046] The received SNR for all N_R received symbol streams, SNR_{rx}, may be defined as:

$$\text{SNR}_{\text{rx}} = \frac{P_{tot} N_R}{\sigma^2} \quad . \qquad \text{Eq (6)}$$

[1047] Combining equations (5) and (6), the effective SNR for the symbol stream recovered in the k-th stage may be expressed as:

$$\text{SNR}_{\text{eff}}(k) = \left(\frac{N_R - N_T + k}{N_T N_R} \right) \text{SNR}_{\text{rx}} \quad . \qquad \text{Eq (7)}$$

The effective SNR formulation shown in equation (7) is based on several assumptions. First, it is assumed that the interference due to each recovered data stream is effectively canceled and does not contribute to the noise and interference observed by the subsequently recovered symbol streams. Second, it is assumed that no (or low) errors propagate from one stage to another. Third, an optimum filter that maximizes SNR is used to obtain each detected symbol stream. Equation (7) also provides the effective SNR in linear unit (i.e., not in log or dB unit).

[1048] As noted above, the transmitted symbol streams may experience different channel conditions and may achieve different SNRs for a given amount of transmit power. If the achieved SNR of each symbol stream is known at the transmitter, then the data rate and coding and modulation scheme for the corresponding data stream may be selected to maximize spectral efficiency while achieving a target packet error rate (PER). However, for some MIMO systems, channel state information indicative of the current channel conditions is not available at the transmitter. In this case, it is not possible to perform adaptive rate control for the data streams.

[1049] Conventionally, in some MIMO systems, data is transmitted over the N_T transmit antennas at the same data rates (i.e., uniform distribution of data rates) when channel state information is not available at the transmitter. At the receiver, the N_R received symbol streams may be processed using the successive cancellation receiver

processing technique. In one conventional scheme, the SNRs of the $(N_T - k + 1)$ detected symbol streams at each stage k are determined, and the detected symbol stream with the highest SNR is recovered in that stage. This transmission scheme with uniform distribution of data rates provides sub-optimal performance.

[1050] Techniques are provided herein to provide improved performance for a MIMO system when channel state information indicative of the current channel conditions is not available at the transmitter. In an aspect, a non-uniform distribution of data rates is used for the transmitted data streams. The data rates may be selected to achieve (1) a given or specified overall spectral efficiency with a lower minimum received SNR or (2) a higher overall spectral efficiency for a given or specified received SNR. A specific scheme for achieving each of the above objectives is provided below. It can be shown that the non-uniform distribution of data rates generally outperforms the conventional uniform distribution of data rates in many situations.

[1051] As shown in equation (7), the effective SNR of each recovered symbol stream is dependent on the particular stage at which it is recovered, as indicated by the factor "k" in the numerator in equation (7). The lowest effective SNR is achieved for the first recovered symbol stream, and the highest effective SNR is achieved for the last recovered symbol stream.

[1052] To achieve improved performance, non-uniform distribution of data rates may be used for the data streams transmitted on different antennas (i.e., different spectral efficiencies may be assigned to different transmit antennas), depending on their effective SNRs. At the receiver, the transmitted data streams may be recovered in an ascending order of data rates. That is, the data stream with the lowest data rate is recovered first, the data stream with the next higher data rate is recovered second, and so on, and the data stream with the highest data rate is recovered last.

[1053] The data rates to be used for the data streams may be determined by taking into account various considerations. First, earlier recovered symbol streams achieve lower effective SNRs, as shown in equation (7), and further suffer from lower diversity order. In fact, the diversity order at stage k may be given as $(N_R - N_T + k)$. Moreover, decoding errors from earlier recovered symbol streams propagate to later recovered symbol streams and can affect the effective SNRs of these subsequently recovered symbol streams. The data rates for earlier recovered symbol streams may thus be selected to achieve high confidence in the recovery of these symbol streams and to

reduce or limit the error propagation (EP) effect on later recovered symbol streams. Second, the later recovered symbol streams may be more vulnerable to errors if they are designated to support larger spectral efficiencies, even though they may be able to achieve higher effective SNRs.

[1054] Various schemes may be implemented to (1) determine the minimum received SNR needed to support a given distribution of data rates (or spectral efficiencies) or, (2) determine the distribution of spectral efficiencies that attains the best performance for a given received SNR. One specific scheme for each of these objectives is described below.

[1055] FIG. 3 is a flow diagram of an embodiment of a process 300 for determining the minimum received SNR needed to support a given set of data rates. This set of data rates is denoted as $\{r_k\}$, for $k=1, 2, \ldots N_T$, and are ordered such that $r_1 \leq r_2 \ldots \leq r_{N_T}$. The data rates in set $\{r_k\}$ are to be used for the N_T data streams to be transmitted from the N_T transmit antennas.

[1056] Initially, the SNR required at the receiver to support each data rate (or spectral efficiency) in set $\{r_k\}$ is determined (step 312). This may be achieved by using a look-up table of required SNR versus spectral efficiency. The required SNR for a given spectral efficiency may be determined (e.g., using computer simulation) based on an assumption that a single data stream is transmitted over a $\{1, N_R\}$ single-input multiple-output (SIMO) channel, and is further determined for a particular target PER (e.g., 1% PER). The required SNR for a data stream with data rate r_k is denoted as $\text{SNR}_{\text{req}}(r_k)$. A set of N_T required SNRs is obtained in step 312 for the N_T data streams.

[1057] The N_T data rates in set $\{r_k\}$ are associated with N_T SNRs required at the receiver to achieve the target PER (e.g., as determined from the look-up table). These N_T data rates are also associated with N_T effective SNRs that may be achieved at the receiver based on a particular received SNR using successive interference cancellation processing at the receiver, as shown in equation (7). The data rates in set $\{r_k\}$ are deemed to be supported if the N_T required SNRs are at or below the corresponding effective SNRs. Visually, the N_T required SNRs may be plotted versus data rates and connected together by a first line, and the N_T effective SNRs may also be plotted versus data rates and connected together by a second line. The data rates in set $\{r_k\}$ are then deemed to be supported if no part of the first line is above the second line.

[1058] The margin for a given data rate may be defined as the difference between the effective SNR and the required SNR for the data rate, i.e., $\text{margin}(k) = \text{SNR}_{\text{eff}}(r_k) - \text{SNR}_{\text{req}}(r_k)$. The data rates in set $\{r_k\}$ may also be deemed to be supported if the margin for each data rate is zero or greater.

[1059] The effective SNRs for the data streams are dependent on the received SNR, and may be derived from the received SNR as shown in equation (7). The minimum received SNR needed to support the N_T data rates in set $\{r_k\}$ is the received SNR that results in the effective SNR of at least one data rate being equal to the required SNR (i.e., zero margin). Depending on the specific data rates included in set $\{r_k\}$, the minimum margin (of zero) may be achieved for any one of the N_T data rates in the set.

[1060] For the first iteration, the minimum margin is assumed to be achieved by the last recovered data stream, and the index variable λ is set to N_T (i.e., $\lambda = N_T$) (step 314). The effective SNR for the λ-th recovered data stream is then set equal to its required SNR (i.e., $\text{SNR}_{\text{eff}}(\lambda) = \text{SNR}_{\text{req}}(\lambda)$) (step 316). The received SNR is next determined based on the effective SNR of $\text{SNR}_{\text{eff}}(\lambda)$ for the λ-th recovered data stream, using equation (7) (step 318). For the first iteration when $\lambda = N_T$, the received SNR may be determined using equation (7) with $k = N_T$, which may then be expressed as:

$$\text{SNR}_{\text{rx}} = N_T \cdot \text{SNR}_{\text{eff}}(N_T) \quad . \qquad \text{Eq (8)}$$

The effective SNR of each remaining data stream is then determined based on the received SNR computed in step 318 and using equation (7), for $k = 1, 2, \ldots N_T - 1$ (step 320). A set of N_T effective SNRs is obtained by step 320 for the N_T data streams.

[1061] The required SNR for each data rate in set $\{r_k\}$ is then compared against the effective SNR for the data rate (step 322). A determination is next made whether or not the data rates in set $\{r_k\}$ are supported by the received SNR determined in step 318 (step 324). In particular, if the required SNR for each of the N_T data rates is less than or equal to the effective SNR for that data rate, then the data rates in set $\{r_k\}$ are deemed to be supported by the received SNR and success is declared (step 326). Otherwise, if any one of the N_T data rates exceeds the effective SNR for the data rate, then the data

rates in set $\{r_k\}$ are deemed to not be supported by the received SNR. In this case, the variable λ is decremented (i.e., $\lambda = \lambda - 1$, so that $\lambda = N_T - 1$ for the second iteration) (step 328). The process then returns to step 316 to determine the set of effective SNRs for the data rates in set $\{r_k\}$ under the assumption that the minimum margin is achieved for the second to last recovered data stream. As many iterations as necessary may be performed until success is declared in step 326. The received SNR determined in step 318 for the iteration that results in the declaration of success is then the minimum received SNR needed to support the data rates in set $\{r_k\}$.

[1062] The process shown in FIG. 3 may also be used to determine whether or not a given set of data rates is supported by a given received SNR. This received SNR may correspond to the operating SNR, SNR_{op}, which may be the average or expected (but not necessarily the instantaneous) received SNR at the receiver. The operating SNR may be determined based on measurements at the receiver and may be periodically provided to the transmitter. Alternatively, the operating SNR may be an estimate of the MIMO channel in which the transmitter is expected to operate. In any case, the received SNR is given or specified for the MIMO system.

[1063] Referring to FIG. 3, to determine whether or not the given set of data rates is supported by the given received SNR, the required SNR for each data rate may be determined initially (step 312). A set of N_T required SNRs is obtained in step 312 for the N_T data streams. Steps 314, 316, and 318 may be skipped, since the received SNR is already given. The effective SNR of each data stream is then determined based on the given received SNR and using equation (7), for $k = 1, 2, \ldots N_T$ (step 320). A set of N_T effective SNRs is obtained in step 320 for the N_T data streams.

[1064] The required SNR for each data rate in set $\{r_k\}$ is then compared against the effective SNR for that data rate (step 322). A determination is next made whether or not the data rates in set $\{r_k\}$ are supported by the received SNR. If the required SNR for each of the N_T data rates is less than or equal to the effective SNR for that data rate, then the data rates in set $\{r_k\}$ are deemed to be supported by the received SNR, and success is declared (step 326). Otherwise, if the required SNR for any one of the N_T data rates exceeds the effective SNR for the data rate, then the data rates in set $\{r_k\}$ are deemed to not be supported by the received SNR, and failure is declared.

[1065] For clarity, an example is described below for a {2, 4} MIMO system with two transmit antennas (i.e., $N_T = 2$) and four received antennas (i.e., $N_R = 4$) and designated to support an overall spectral efficiency of 3 bits per second per Hertz (bps/Hz). For this example, two sets of data rates are evaluated. The first set includes data rates corresponding to 1 bps/Hz and 2 bps/Hz, and the second set includes data rates corresponding to 4/3 bps/Hz and 5/3 bps/Hz. The performance of each rate set is determined (e.g., based on the process shown in FIG. 3) and compared against one another.

[1066] FIG. 4 shows plots of PER versus SNR for a {1, 4} MIMO system for spectral efficiencies of 1 bps/Hz, 4/3 bps/Hz, 5/3 bps/Hz, and 2 bps/Hz. These plots may be generated by computer simulation or some other means, as is known in the art. A MIMO system is typically designated to operate at a particular target PER. In this case, the SNR required to achieve the target PER for each spectral efficiency may be determined and stored in a look-up table. For example, if the target PER is 1%, then values of -2.0 dB, 0.4 dB, 3.1 dB, and 3.2 dB may be stored in the look-up table for spectral efficiencies of 1, 4/3, 5/3, and 2 bps/Hz, respectively.

[1067] For the first rate set, the required SNRs for data streams 1 and 2 with spectral efficiencies of 1 and 2 bps/Hz, respectively, may be determined (step 312 in FIG. 3) using plots 412 and 418 in FIG. 4, as follows:

$\mathrm{SNR}_{\mathrm{req}}(1) = -2.0\,\mathrm{dB}$, for data stream 1 with spectral efficiency of 1 bps/Hz, and

$\mathrm{SNR}_{\mathrm{req}}(2) = 3.2\,\mathrm{dB}$, for data stream 2 with spectral efficiency of 2 bps/Hz.

The effective SNR of data stream 2 (which is recovered last and under the assumption that the interference from data stream 1 was effectively cancelled) is then set to its required SNR (step 316), as follows:

$$\mathrm{SNR}_{\mathrm{eff}}(2) = \mathrm{SNR}_{\mathrm{req}}(2) = 3.2\,\mathrm{dB}\ .$$

The received SNR is then determined based on equation (8) (step 318), as follows:

$$\mathrm{SNR}_{\mathrm{rx}} = 2 \cdot \mathrm{SNR}_{\mathrm{req}}(2)\ , \qquad \text{for linear unit, or}$$

$$\mathrm{SNR}_{\mathrm{rx}} = \mathrm{SNR}_{\mathrm{req}}(2) + 3.0\,\mathrm{dB} = 6.2\,\mathrm{dB}\ , \qquad \text{for log unit.}$$

[1068] The effective SNR of each remaining data stream (i.e., data stream 1) is next determined based on equation (7) (step 320), as follows:

$$\mathrm{SNR}_{\mathrm{eff}}(1) = 3/8 \cdot \mathrm{SNR}_{\mathrm{rx}}\ , \qquad \text{for linear unit, or}$$

$$\mathrm{SNR}_{\mathrm{eff}}(1) = \mathrm{SNR}_{\mathrm{rx}} - 4.3\,\mathrm{dB} = 1.9\,\mathrm{dB}\ , \qquad \text{for log unit.}$$

[1069] The effective and required SNRs for each data rate in the first rate set are given in columns 2 and 3 in Table 1. The margin for each data rate is also determined and given in the last row in Table 1.

Table 1

	First rate set		Second rate set		Unit
Data stream	1	2	1	2	
Spectral efficiency	1	2	4/3	5/3	bps/Hz
$\mathrm{SNR}_{\mathrm{eff}}$	1.9	3.2	1.8	3.1	dB
$\mathrm{SNR}_{\mathrm{req}}$	-2.0	3.2	0.4	3.1	dB
margin	3.9	0.0	1.4	0.0	dB

[1070] The required SNRs for data stream 1 and 2 are then compared against the effective SNRs for these data streams (step 322). Since $\mathrm{SNR}_{\mathrm{req}}(2) = \mathrm{SNR}_{\mathrm{eff}}(2)$ and $\mathrm{SNR}_{\mathrm{req}}(1) < \mathrm{SNR}_{\mathrm{eff}}(1)$, this set of data rates is supported by a minimum received SNR of 6.2 dB.

[1071] Since the first rate set is deemed to be supported by the first iteration through the process shown in FIG. 3, no additional iterations need to be performed. However, had this first rate set not been supported by a received SNR of 6.2 dB (e.g., if the required SNR for data stream 1 turned out to be greater than 1.9 dB), then another iteration would be performed whereby the received SNR is determined based on $\mathrm{SNR}_{\mathrm{req}}(1)$ and would be greater than 6.2 dB.

[1072] For the second rate set, the required SNRs for data streams 1 and 2 with spectral efficiencies of 4/3 and 5/3 bps/Hz, respectively, may be determined using plots 414 and 416 in FIG. 4, as follows:

$\mathrm{SNR}_{\mathrm{req}}(1) = 0.4\,\mathrm{dB}$, for data stream 1 with spectral efficiency of 4/3 bps/Hz, and

$$\mathrm{SNR}_{\mathrm{req}}(2) = 3.1\,\mathrm{dB}\,, \quad \text{for data stream 2 with spectral efficiency of 5/3 bps/Hz.}$$

The effective SNR of data stream 2 is then set to its required SNR. The received SNR is then determined based on equation (8), as follows:

$$\mathrm{SNR}_{\mathrm{rx}} = \mathrm{SNR}_{\mathrm{req}}(2) + 3.0\,\mathrm{dB} = 6.1\,\mathrm{dB}\,, \qquad \text{for log unit.}$$

[1073] The effective SNR of each remaining data rate (i.e., data rate 1) is next determined based on equation (7), as follows:

$$\mathrm{SNR}_{\mathrm{eff}}(1) = \mathrm{SNR}_{\mathrm{rx}} - 4.3\,\mathrm{dB} = 1.8\,\mathrm{dB}\,, \qquad \text{for log unit.}$$

[1074] The effective and required SNRs for each data rate in the second rate set are given in columns 4 and 5 in Table 1.

[1075] The effective SNRs of data streams 1 and 2 are then compared against their required SNRs. Again, since $\mathrm{SNR}_{\mathrm{req}}(2) = \mathrm{SNR}_{\mathrm{eff}}(2)$ and $\mathrm{SNR}_{\mathrm{req}}(1) < \mathrm{SNR}_{\mathrm{eff}}(1)$, this set of data rates is supported by a minimum received SNR of 6.1 dB.

[1076] The above description is for a "vertical" successive interference cancellation scheme whereby one data stream is transmitted from each transmit antenna and, at the receiver, one data stream is recovered at each stage of the successive interference cancellation receiver by processing the stream from one transmit antenna. The plots in FIG. 4 and the look-up table are derived for this vertical scheme.

[1077] The techniques described herein may also be used for a "diagonal" successive interference cancellation scheme whereby each data stream is transmitted from multiple (e.g., all N_T) transmit antennas (and possibly across multiple frequency bins). At the receiver, the symbols from one transmit antenna may be detected at each stage of the successive interference cancellation receiver, and each data stream may then be recovered from the symbols detected from multiple stages. For the diagonal scheme, another set of plots and another look-up table may be derived and used. The techniques described herein may also be used for other ordering schemes, and this is within the scope of the invention.

[1078] For the above example, it can be shown that, for the diagonal successive interference cancellation scheme, the minimum received SNR needed to support a uniform distribution of data rates (i.e., spectral efficiency of 1.5 bps/Hz on each of the

two data streams) is approximately 0.6 dB higher than that needed for the second rate set (i.e., spectral efficiencies of 4/3 and 5/3). This gain is achieved without severely complicating the system design.

[1079] In order to reduce the minimum received SNR needed to achieve the target PER for a given overall spectral efficiency, the last recovered data stream may be assigned with the smallest possible spectral efficiency that does not violate the no error propagation condition for any of the prior recovered data streams. If the spectral efficiency of the last recovered data stream is reduced, then the spectral efficiency of one or more prior recovered data streams needs to be increased accordingly to achieve the given overall spectral efficiency. The increased spectral efficiency for the earlier recovered data streams would then result in higher required SNRs. If the spectral efficiency of any one of the earlier recovered data streams is increased too high, then the minimum received SNR is determined by the required SNR for this data stream and not by the last recovered data stream (which is the case for the uniform distribution of data rates).

[1080] In the above example, the second rate set needs a smaller received SNR because the later recovered data stream 2 is assigned a smaller spectral efficiency that does not violate the no error propagation condition for the first recovered data stream 1. For the first rate set, the spectral efficiency assigned to data stream 1 is too conservative so that, while it assures no error propagation, it also hurts the overall performance by forcing a higher spectral efficiency to be assigned to data stream 2. In comparison, the second rate set assigns a more realistic spectral efficiency to data stream 1 that still assures no error propagation (albeit with less confidence in comparison to the first rate set). As shown in Table 1, the margin for data stream 1 for the first rate set is 3.9 dB while the margin for data stream 1 for the second rate set is 1.4 dB.

[1081] The techniques described herein may also be used to determine a set of data rates that maximizes the overall spectral efficiency for a given received SNR (which may be the operating SNR for the MIMO system). In this case, a set of effective SNRs may be initially determined for the N_T data streams based on the given received SNR and using equation (7). For each effective SNR in the set, the highest spectral efficiency that may be supported by this effective SNR for the target PER is then determined. This may be achieved by using another look-up table that stores values for spectral efficiency versus effective SNR. A set of N_T spectral efficiencies is obtained

for the set of N_T effective SNRs. A set of data rates corresponding to this set of N_T spectral efficiencies is then determined and may be used for the N_T data streams. This rate set maximizes the overall spectral efficiency for the given received SNR.

[1082] In the description above, the effective SNRs of the data streams are determined based on the received SNR and using equation (7). This equation includes various assumptions, as noted above, which are generally true (to a large extent) for typically MIMO systems. Moreover, equation (7) is also derived based on the use of successive interference cancellation processing at the receiver. A different equation or a look-up table may also be used to determine the effective SNRs of the data streams for different operating conditions and/or different receiver processing techniques, and this is within the scope of the invention.

[1083] For simplicity, the data rate determination has been described specifically for a MIMO system. These techniques may also be used for other multi-channel communication systems.

[1084] A wideband MIMO system may experience frequency selective fading, which is characterized by different amounts of attenuation across the system bandwidth. This frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. This distortion degrades performance by impacting the ability to correctly detect the received symbols.

[1085] OFDM may be used to combat ISI and/or for some other considerations. An OFDM system effectively partitions the overall system bandwidth into a number of (N_F) frequency subchannels, which may also be referred to as subbands or frequency bins. Each frequency subchannel is associated with a respective subcarrier on which data may be modulated. The frequency subchannels of the OFDM system may also experience frequency selective fading, depending on the characteristics (e.g., the multipath profile) of the propagation path between the transmit and receive antennas. Using OFDM, the ISI due to frequency selective fading may be combated by repeating a portion of each OFDM symbol (i.e., appending a cyclic prefix to each OFDM symbol), as is known in the art.

[1086] For a MIMO system that utilizes OFDM (i.e., a MIMO-OFDM system), N_F frequency subchannels are available on each of the N_S spatial subchannels for data transmission. Each frequency subchannel of each spatial subchannel may be referred to

as a transmission channel, and $N_F \cdot N_S$ transmission channels are available for data transmission between the N_T transmit antennas and N_R receive antennas. The data rate determination described above may be performed for the set of N_T transmit antennas, similar to that described above for the MIMO system. Alternatively, the data rate determination may be performed independently for the set of N_T transmit antennas for each of the N_F frequency subchannels

Transmitter System

[1087] FIG. 5 is a block diagram of a transmitter unit 500, which is an embodiment of the transmitter portion of transmitter system 110 in FIG. 1. In this embodiment, a separate data rate and coding and modulation scheme may be used for each of the N_T data streams to be transmitted on the N_T transmit antennas (i.e., separate coding and modulation on a per-antenna basis). The specific data rate and coding and modulation schemes to be used for each transmit antenna may be determined based on controls provided by controller 130, and the data rates may be determined as described above.

[1088] Transmitter unit 500 includes (1) a TX data processor 114a that receives, codes, and modulates each data stream in accordance with a separate coding and modulation scheme to provide modulation symbols and (2) a TX MIMO processor 120a that may further process the modulation symbols to provide transmission symbols if OFDM is employed. TX data processor 114a and TX MIMO processor 120a are one embodiment of TX data processor 114 and TX MIMO processor 120, respectively, in FIG. 1.

[1089] In the specific embodiment shown in FIG. 5, TX data processor 114a includes a demultiplexer 510, N_T encoders 512a through 512t, N_T channel interleavers 514a through 514t, and N_T symbol mapping elements 516a through 516t, (i.e., one set of encoder, channel interleaver, and symbol mapping element for each transmit antenna). Demultiplexer 510 demultiplexes the traffic data (i.e., the information bits) into N_T data streams for the N_T transmit antennas to be used for data transmission. The N_T data streams may be associated with different data rates, as determined by the rate control. Each data stream is provided to a respective encoder 512.

[1090] Each encoder 512 receives and codes a respective data stream based on the specific coding scheme selected for that data stream to provide coded bits. The coding increases the reliability of the data transmission. The coding scheme may include any

combination of cyclic redundancy check (CRC) coding, convolutional coding, Turbo coding, block coding, and so on. The coded bits from each encoder 512 are then provided to a respective channel interleaver 514, which interleaves the coded bits based on a particular interleaving scheme. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average SNR for the transmission channels used for the data stream, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1091] The coded and interleaved bits from each channel interleaver 514 are provided to a respective symbol mapping element 516, which maps these bits to form modulation symbols. The particular modulation scheme to be implemented by each symbol mapping element 516 is determined by the modulation control provided by controller 130. Each symbol mapping element 516 groups sets of q_j coded and interleaved bits to form non-binary symbols, and further maps each non-binary symbol to a specific point in a signal constellation corresponding to the selected modulation scheme (e.g., QPSK, M-PSK, M-QAM, or some other modulation scheme). Each mapped signal point corresponds to an M_j-ary modulation symbol, where M_j corresponds to the specific modulation scheme selected for the j-th transmit antenna and $\mathrm{M}_j = 2^{q_j}$. Symbol mapping elements 516a through 516t then provide N_T streams of modulation symbols.

[1092] In the specific embodiment shown in FIG. 5, TX MIMO processor 120a includes N_T OFDM modulators, with each OFDM modulator including an inverse Fourier transform (IFFT) unit 522 and a cyclic prefix generator 524. Each IFFT 522 receives a respective modulation symbol stream from a corresponding symbol mapping element 516. Each IFFT 522 groups sets of N_F modulation symbols to form corresponding modulation symbol vectors, and converts each modulation symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using the inverse fast Fourier transform. IFFT 522 may be designed to perform the inverse transform on any number of frequency subchannels (e.g., 8, 16, 32, … , N_F, …). For each OFDM symbol, cyclic prefix generator 524 repeats a portion of the OFDM symbol to form a corresponding transmission symbol. The cyclic prefix ensures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects such as channel dispersion caused by frequency selective fading. Cyclic prefix generator 524 then

provides a stream of transmission symbols to an associated transmitter 122. If OFDM is not employed, then TX MIMO processor 120a simply provides the modulation symbol stream from each symbol mapping element 516 to the associated transmitter 122.

[1093] Each transmitter 122 receives and processes a respective modulation symbol stream (for MIMO without OFDM) or transmission symbol stream (for MIMO with OFDM) to generate a modulated signal, which is then transmitted from the associated antenna 124.

[1094] Other designs for the transmitter unit may also be implemented and are within the scope of the invention.

[1095] The coding and modulation for MIMO systems with and without OFDM are described in further detail in the following U.S. patent applications:

- U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001;
- U.S. Patent Application Serial No. 09/854,235, entitled "Method and Apparatus for Processing Data in a Multiple-Input Multiple-Output (MIMO) Communication System Utilizing Channel State Information," filed May 11, 2001;
- U.S. Patent Application Serial Nos. 09/826,481 and 09/956,449, both entitled "Method and Apparatus for Utilizing Channel State Information in a Wireless Communication System," respectively filed March 23, 2001 and September 18, 2001;
- U.S. Patent Application Serial No. 09/776,075, entitled "Coding Scheme for a Wireless Communication System," filed February 1, 2001; and
- U.S. Patent Application Serial No. 09/532,492, entitled "High Efficiency, High Performance Communications System Employing Multi-Carrier Modulation," filed March 30, 2000.

These applications are all assigned to the assignee of the present application and incorporated herein by reference. Application Serial No. 09/776,075 describes a coding scheme whereby different rates may be achieved by coding the data with the same base code (e.g., a convolutional or Turbo code) and adjusting the puncturing to achieve the

desired rate. Other coding and modulation schemes may also be used, and this is within the scope of the invention.

Receiver System

[1096] FIG. 6 is a block diagram of a RX MIMO/data processor 160a capable of implementing the successive cancellation receiver processing technique. RX MIMO/data processor 160a is one embodiment of RX MIMO/data processor 160 in FIG. 1. The signals transmitted from N_T transmit antennas are received by each of N_R antennas 152a through 152r and routed to a respective receiver 154. Each receiver 154 conditions (e.g., filters, amplifies, and downconverts) a respective received signal and digitizes the conditioned signal to provide a corresponding stream of data samples.

[1097] For MIMO without OFDM, the data samples are representative of the received symbols. Each receiver 154 would then provide to RX MIMO/data processor 160a a respective received symbol stream, which includes a received symbol for each symbol period.

[1098] For MIMO with OFDM, each receiver 154 further includes a cyclic prefix removal element and an FFT processor (both of which are not shown in FIG. 6 for simplicity). The cyclic prefix removal element removes the cyclic prefix, which has been inserted at the transmitter system for each transmission symbol, to provide a corresponding received OFDM symbol. The FFT processor then transforms each received OFDM symbol to provide a vector of N_F received symbols for the N_F frequency subchannels for that symbol period. N_R received symbol vector streams are then provided by N_R receivers 154 to RX MIMO/data processor 160a.

[1099] For MIMO with OFDM, RX MIMO/data processor 160a may demultiplex the N_R received symbol vector streams into N_F groups of N_R received symbol streams, one group for each frequency subchannel, with each group including N_R streams of received symbols for one frequency subchannel. RX MIMO/data processor 160a may then process each group of N_R received symbol streams in similar manner as for the N_R received symbol streams for MIMO without OFDM. RX MIMO/data processor 160a may also process the received symbols for MIMO with OFDM based on some other ordering scheme, as is known in the art. In any case, RX MIMO/data processor 160a processes the N_R received symbol streams (for MIMO without OFDM) or each group of N_R received symbol streams (for MIMO with OFDM).

[1100] In the embodiment shown in FIG. 6, RX MIMO/data processor 160a includes a number of successive (i.e., cascaded) receiver processing stages 610a through 610n, one stage for each of the transmitted data streams to be recovered. Each receiver processing stage 610 (except for the last stage 610n) includes a spatial processor 620, an RX data processor 630, and an interference canceller 640. The last stage 610n includes only spatial processor 620n and RX data processor 630n.

[1101] For the first stage 610a, spatial processor 620a receives and processes the N_R received symbol streams (denoted as the vector $\underline{\mathbf{y}}^1$) from receivers 154a through 154r based on a particular spatial or space-time receiver processing technique to provide (up to) N_T detected symbol streams (denoted as the vector $\hat{\underline{\mathbf{x}}}^1$). For MIMO with OFDM, the N_R received symbol streams comprise the received symbols for one frequency subchannel. The detected symbol stream corresponding to the lowest data rate, $\hat{x}_1$, is selected and provided to RX data processor 630a. Processor 630a further processes (e.g., demodulates, deinterleaves, and decodes) the detected symbol stream, $\hat{x}_1$, selected for the first stage to provide a decoded data stream. Spatial processor 620a further provides an estimate of the channel response matrix $\underline{\mathbf{H}}$, which is used to perform the spatial or space-time processing for all stages.

[1102] For the first stage 610a, interference canceller 640a also receives the N_R received symbol streams from receivers 154 (i.e., the vector $\underline{\mathbf{y}}^1$). Interference canceller 640a further receives the decoded data stream from RX data processor 630a and performs the processing (e.g., encoding, interleaving, modulation, channel response, and so on) to derive N_R remodulated symbol streams (denoted as the vector $\underline{\mathbf{i}}^1$) that are estimates of the interference components due to the just-recovered data stream. The remodulated symbol streams are then subtracted from the first stage's input symbol streams to derive N_R modified symbol streams (denoted as the vector $\underline{\mathbf{y}}^2$), which include all but the subtracted (i.e., cancelled) interference components. The N_R modified symbol streams are then provided to the next stage.

[1103] For each of the second through last stages 610b through 610n, the spatial processor for that stage receives and processes the N_R modified symbol streams from the interference canceller in the preceding stage to derive the detected symbol streams for that stage. The detected symbol stream corresponding to the lowest data rate at that

stage is selected and processed by the RX data processor to provide the decoded data stream for that stage. For each of the second through second-to-last stages, the interference canceller in that stage receives the N_R modified symbol streams from the interference canceller in the preceding stage and the decoded data stream from the RX data processor within the same stage, derives N_R remodulated symbol streams, and provides N_R modified symbol streams for the next stage.

[1104] The successive cancellation receiver processing technique is described in further detail in the aforementioned U.S. Patent Application Serial Nos. 09/993,087 and 09/854,235.

[1105] The spatial processor 620 in each stage implements a particular spatial or space-time receiver processing technique. The specific receiver processing technique to be used is typically dependent on the characteristics of the MIMO channel, which may be characterized as either non-dispersive or dispersive. A non-dispersive MIMO channel experiences flat fading (i.e., approximately equal amount of attenuation across the system bandwidth), and a dispersive MIMO channel experiences frequency-selective fading (e.g., different amounts of attenuation across the system bandwidth).

[1106] For a non-dispersive MIMO channel, spatial receiver processing techniques may be used to process the received signals to provide the detected symbol streams. These spatial receiver processing techniques include a channel correlation matrix inversion (CCMI) technique (which is also referred to as a zero-forcing technique) and a minimum mean square error (MMSE) technique. Other spatial receiver processing techniques may also be used and are within the scope of the invention.

[1107] For a dispersive MIMO channel, time dispersion in the channel introduces inter-symbol interference (ISI). To improve performance, a receiver attempting to recover a particular transmitted data stream would need to ameliorate both the interference (or "crosstalk") from the other transmitted data streams as well as the ISI from all data streams. To combat both crosstalk and ISI, space-time receiver processing techniques may be used to process the received signals to provide the detected symbol streams. These space-time receiver processing techniques include a MMSE linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum-likelihood sequence estimator (MLSE), and so on.

[1108] The CCMI, MMSE, MMSE-LE, and DFE techniques are described in detail in the aforementioned U.S. Patent Application Serial Nos. 09/993,087, 09/854,235, 09/826,481, and 09/956,44.

[1109] The data rate determination and data transmission techniques described herein may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements used to determinate data rates at the transmitter and the data transmission at the transmitter/receiver may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1110] For a software implementation, certain aspects of the data rate determination and the processing at the transmitter/receiver may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 132 in FIG. 1) and executed by a processor (e.g., controller 130). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[1111] Headings are included herein for reference and to aid in locating certain sections. These headings are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[1112] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1113] **WHAT IS CLAIMED IS:**

CLAIMS

1. A method for determining data rates for a plurality of data streams to be transmitted via a plurality of transmission channels in a multi-channel communication system, comprising:

determining a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for the plurality of data streams, wherein at least two of the data rates are unequal;

determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

comparing the required SNR for each data stream against the effective SNR for the data stream; and

determining whether or not the plurality of data rates are supported based on results of the comparing.

2. The method of claim 1, wherein the plurality of data streams are transmitted over a plurality of transmit antennas in a multiple-input multiple-output (MIMO) communication system.

3. The method of claim 2, wherein each data stream is transmitted over a respective transmit antenna, and wherein the effective SNR for each data stream is determined based on full transmit power being used for the data stream.

4. The method of claim 1, wherein the effective SNR for each data stream is further determined based on a received SNR indicative of an operating condition of the plurality of transmission channels.

5. The method of claim 4, wherein the received SNR is determined based on the required SNR for one of the plurality of data streams.

6. The method of claim 4, wherein the received SNR is specified for the communication system.

7. The method of claim 4, wherein the received SNR is estimated at the receiver.

8. The method of claim 4, wherein the successive interference cancellation processing recovers one data stream at each stage, and wherein the effective SNR for each recovered data stream is estimated as

$$\mathrm{SNR}_{\mathrm{eff}}(k) = \left(\frac{N_R - N_T + k}{N_T N_R} \right) \mathrm{SNR}_{\mathrm{rx}} \ , \qquad \text{Eq (9)}$$

where $\mathrm{SNR}_{\mathrm{eff}}(k)$ is the effective SNR for the data stream recovered in stage *k*,

$\mathrm{SNR}_{\mathrm{rx}}$ is the received SNR,

N_T is the number of transmit antennas used for data transmission, and

N_R is the number of receive antennas.

9. The method of claim 4, further comprising:

evaluating a plurality of sets of data rates; and

selecting a rate set associated with a minimum received SNR for use for the plurality of data streams.

10. The method of claim 9, wherein the data rates in each rate set are selected to achieve a specified overall spectral efficiency.

11. The method of claim 1, wherein the required SNR for each data rate is determined based on a look-up table.

12. The method of claim 1, wherein the plurality of data rates are deemed to be supported if the required SNR for each data rate is less than or equal to the effective SNR for the data rate.

13. The method of claim 1, wherein the communication system implements orthogonal frequency division multiplexing (OFDM).

14. A method for determining data rates for a plurality of data streams to be transmitted over a plurality of transmit antennas in a multiple-input multiple-output (MIMO) communication system, comprising:

determining an operating signal-to-noise-and-interference ratio (SNR) indicative of an operating condition of the MIMO system;

determining a required SNR for each of a plurality of data rates to be used for the plurality of data streams, wherein at least two of the data rates are unequal and wherein the plurality of data rates are selected to achieve a specified overall spectral efficiency;

determining an effective SNR for each of the plurality of data streams based on the operating SNR and successive interference cancellation processing technique at a receiver to recover the plurality of data streams;

comparing the required SNR for each data stream against the effective SNR for the data stream; and

determining whether or not the plurality of data rates are supported based on results of the comparing.

15. A method for determining data rates for a plurality of data streams to be transmitted via a plurality of transmission channels in a multi-channel communication system, comprising:

determining a received SNR indicative of an operating condition of the plurality of transmission channels;

determining an effective SNR for each of the plurality of data streams based on the received SNR and successive interference cancellation processing at a receiver to recover the plurality of data streams; and

determining a data rate for each data stream based on the effective SNR for the data stream, wherein at least two of the data rates are unequal.

16. The method of claim 15, wherein the data rate for each data stream is determined such that a required SNR for the data stream is less than or equal to the effective SNR for the data stream.

17. The method of claim 15, wherein the received SNR is specified for the communication system.

18. The method of claim 15, wherein each data stream is transmitted over a respective transmit antenna in a multiple-input multiple-output (MIMO) communication system.

19. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

determine a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for a plurality of data streams to be transmitted via a plurality of transmission channels in a multi-channel communication system, wherein at least two of the data rates are unequal;

determine an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

compare the required SNR for each data stream against the effective SNR for the data stream; and

determine whether or not the plurality of data rates are supported based on results of the comparison.

20. An apparatus in a multi-channel communication system, comprising:

means for determining a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for a plurality of data streams to be transmitted via a plurality of transmission channels, wherein at least two of the data rates are unequal;

means for determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

means for comparing the required SNR for each data stream against the effective SNR for the data stream; and

means for determining whether or not the plurality of data rates are supported based on results of the comparing.

21. The apparatus of claim 20, further comprising:

means for evaluating a plurality of sets of data rates; and

means for selecting a rate set associated with a minimum received SNR for use for the plurality of data streams.

22. The apparatus of claim 20, wherein the multi-channel communication system is a multiple-input multiple-output (MIMO) communication system.

23. The apparatus of claim 22, wherein the MIMO system implements orthogonal frequency division multiplexing (OFDM).

24. A base station comprising the apparatus of claim 20.

25. A wireless terminal comprising the apparatus of claim 20.

26. A transmitter unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a controller operative to determine a plurality of data rates for a plurality of data streams to be transmitted over a plurality of transmit antennas by

determining a required signal-to-noise-and-interference ratio (SNR) for each of the plurality of data rates, wherein at least two of the data rates are unequal,

determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing technique at a receiver to recover the plurality of data streams,

comparing the required SNR for each data stream against the effective SNR for the data stream, and

determining whether or not the plurality of data rates are supported based on results of the comparing;

a transmit (TX) data processor operative to process each data stream with the determined data rate to provide a respective symbol stream; and

one or more transmitters operative to process a plurality of symbol streams for the plurality of data streams to provide a plurality of modulated signals suitable for transmission over the plurality of transmit antennas.

27. The transmitter unit of claim 26, wherein the controller is further operative to determine the data rates for the plurality of data streams by

evaluating a plurality of sets of data rates, and

selecting a rate set associated with a minimum received SNR.

28. A base station comprising the transmitter unit of claim 26.

29. A wireless terminal comprising the transmitter unit of claim 26.

30. A transmitter apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for determining a required signal-to-noise-and-interference ratio (SNR) for each of a plurality of data rates to be used for a plurality of data streams to be transmitted over a plurality of transmit antennas in the MIMO system, wherein at least two of the data rates are unequal;

means for determining an effective SNR for each of the plurality of data streams based in part on successive interference cancellation processing at a receiver to recover the plurality of data streams;

means for comparing the required SNR for each data stream against the effective SNR for the data stream;

means for determining whether or not the plurality of data rates are supported based on results of the comparison;

means for processing each data stream to provide a respective symbol stream; and

means for processing a plurality of symbol streams for the plurality of data streams to provide a plurality of modulated signals suitable for transmission over the plurality of transmit antennas.

31. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a receive (RX) MIMO processor operative to receive and process a plurality of received symbol streams using successive interference cancellation processing to provide a plurality of detected symbol streams for a plurality of transmitted data streams, one detected data stream for each stage of the successive interference cancellation processing; and

a RX data processor operative to process each detected symbol stream to provide a corresponding decoded data stream, and

wherein data rates for the plurality of transmitted data streams are determined by determining a received signal-to-noise-and-interference ratio (SNR) indicative of an operating condition of the communication system, determining an effective SNR for each of the plurality of data streams based on the received SNR and the successive interference cancellation processing, and determining the data rate for each data stream based on the effective SNR, and wherein at least two of the data rates are unequal.

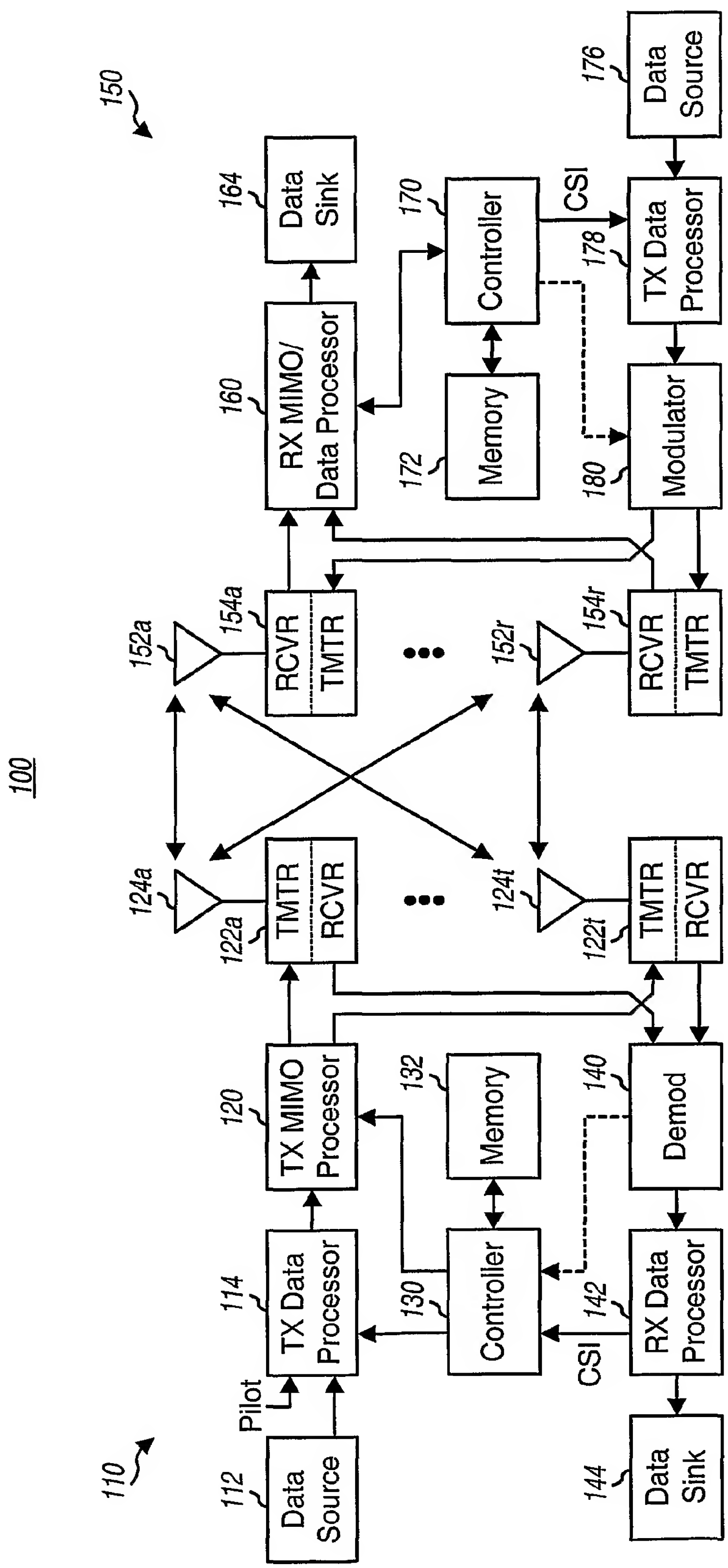

100
110
150
112
Data Source
Pilot
114
TX Data Processor
120
TX MIMO Processor
130
Controller
132
Memory
140
Demod
142
RX Data Processor
CSI
144
Data Sink
122a
TMTR
RCVR
124a
122t
124t
152a
152r
154a
RCVR
TMTR
154r
160
RX MIMO/ Data Processor
164
Data Sink
170
Controller
172
Memory
CSI
176
Data Source
178
TX Data Processor
180
Modulator

FIG. 1

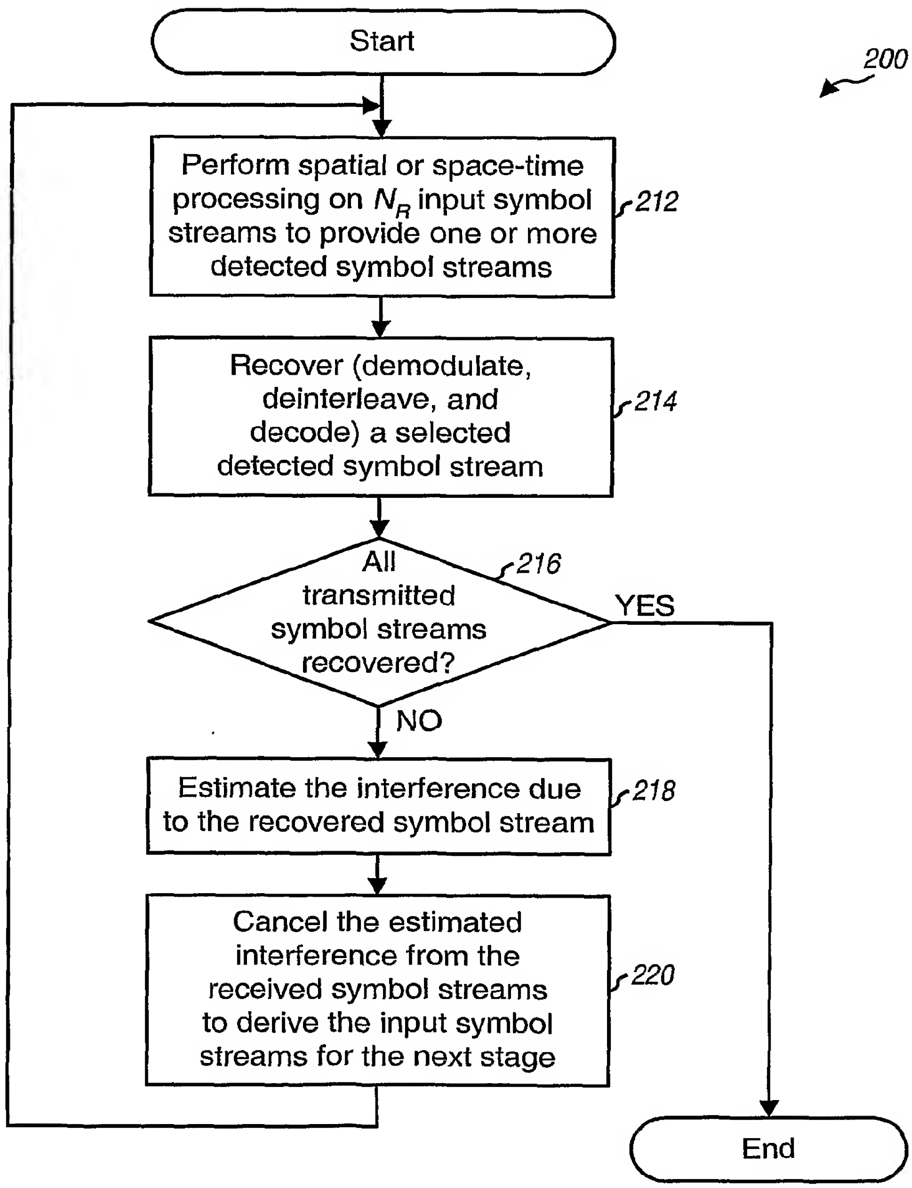

Start
200
Perform spatial or space-time processing on N_R input symbol streams to provide one or more detected symbol streams
212
Recover (demodulate, deinterleave, and decode) a selected detected symbol stream
214
All transmitted symbol streams recovered?
216
YES
NO
Estimate the interference due to the recovered symbol stream
218
Cancel the estimated interference from the received symbol streams to derive the input symbol streams for the next stage
220
End

FIG. 2

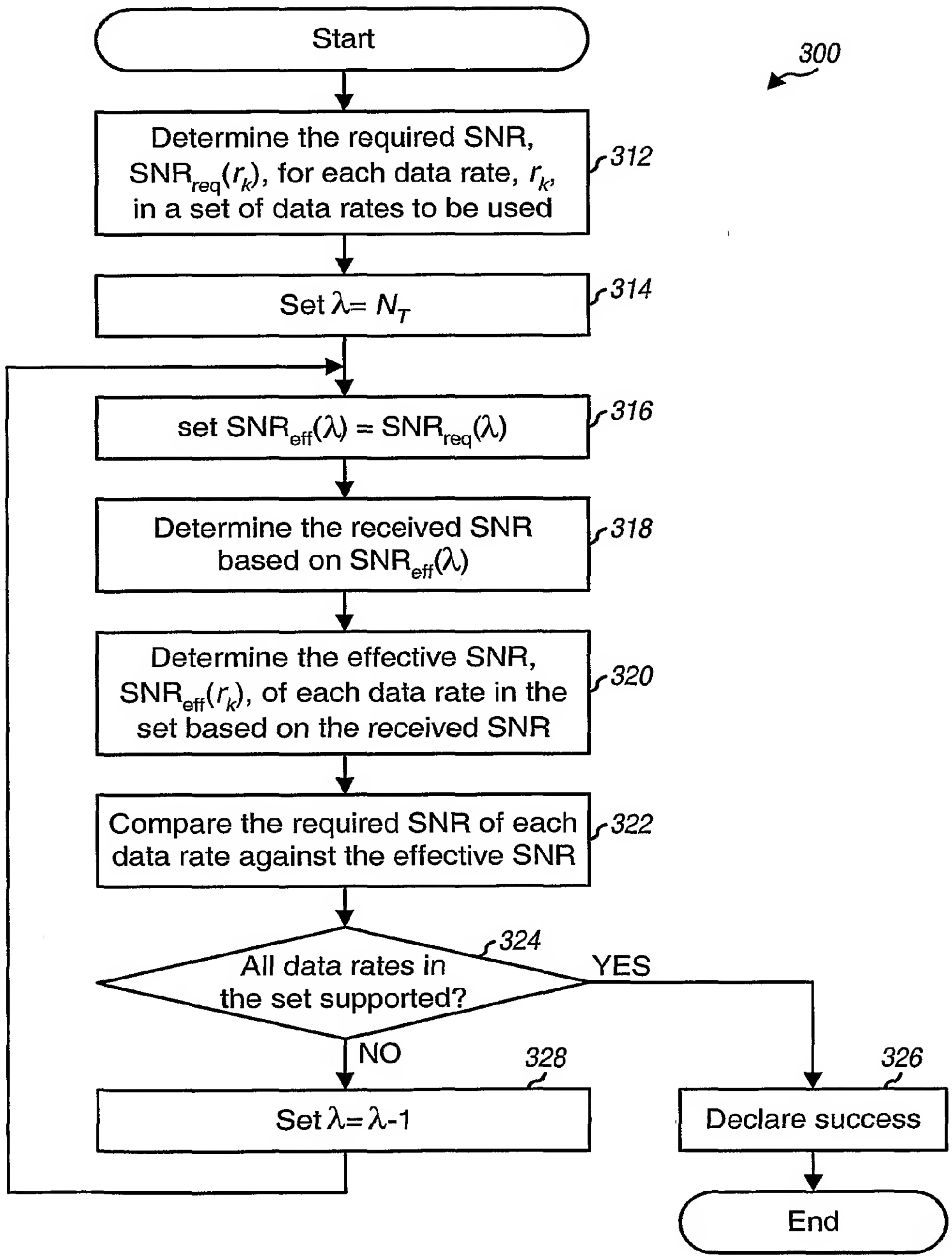

300
Start
Determine the required SNR, $SNR_{req}(r_k)$, for each data rate, r_k, in a set of data rates to be used
312
Set $\lambda = N_T$
314
set $SNR_{eff}(\lambda) = SNR_{req}(\lambda)$
316
Determine the received SNR based on $SNR_{eff}(\lambda)$
318
Determine the effective SNR, $SNR_{eff}(r_k)$, of each data rate in the set based on the received SNR
320
Compare the required SNR of each data rate against the effective SNR
322
All data rates in the set supported?
324
YES
NO
Set $\lambda = \lambda - 1$
328
Declare success
326
End

FIG. 3

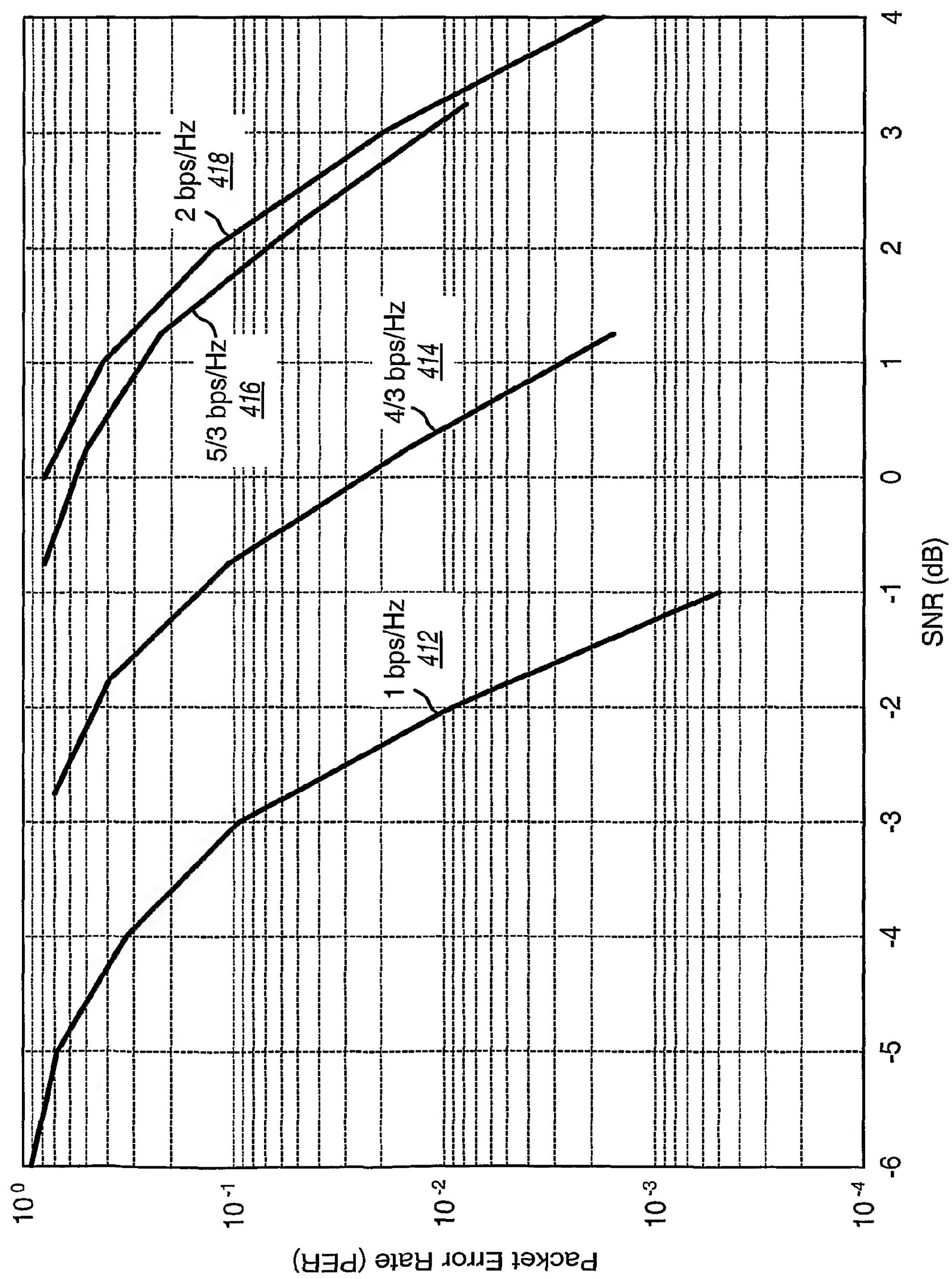

Packet Error Rate (PER)
10^{0}
10^{-1}
10^{-2}
10^{-3}
10^{-4}
SNR (dB)
-6
-5
-4
-3
-2
-1
0
1
2
3
4
1 bps/Hz
412
4/3 bps/Hz
414
5/3 bps/Hz
416
2 bps/Hz
418

FIG. 4

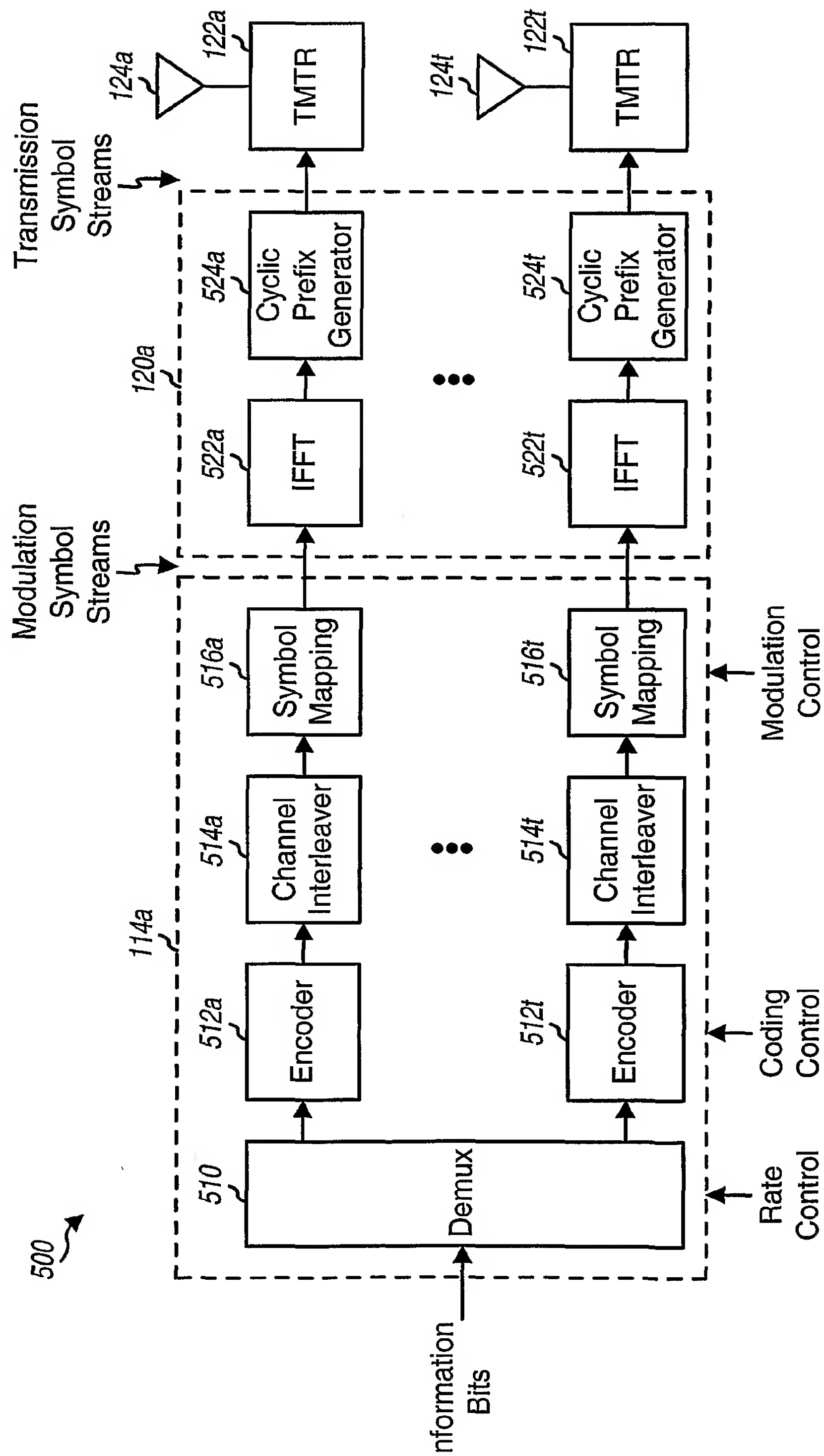
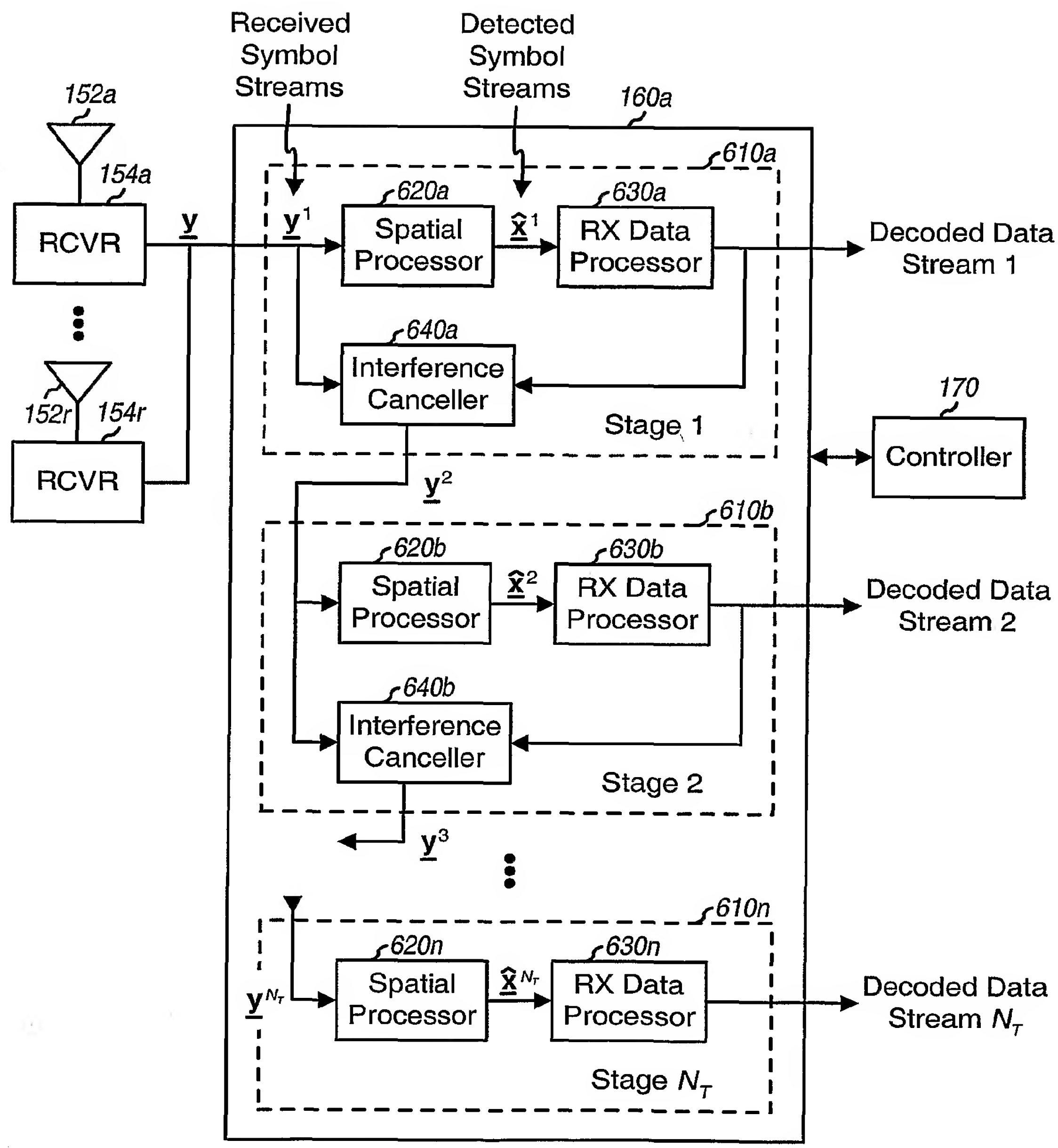

500
Information Bits
510
Demux
114a
512a
Encoder
512t
Encoder
514a
Channel Interleaver
514t
Channel Interleaver
516a
Symbol Mapping
516t
Symbol Mapping
Rate Control
Coding Control
Modulation Control
Modulation Symbol Streams
120a
522a
IFFT
522t
IFFT
524a
Cyclic Prefix Generator
524t
Cyclic Prefix Generator
Transmission Symbol Streams
122a
TMTR
122t
TMTR
124a
124t

FIG. 5



Received Symbol Streams
Detected Symbol Streams
152a
154a
RCVR
y
152r
154r
RCVR
160a
610a
y1
620a
Spatial Processor
x̂1
630a
RX Data Processor
640a
Interference Canceller
Stage 1
y2
610b
620b
Spatial Processor
x̂2
630b
RX Data Processor
640b
Interference Canceller
Stage 2
y3
610n
yNT
620n
Spatial Processor
x̂NT
630n
RX Data Processor
Stage NT
Decoded Data Stream 1
Decoded Data Stream 2
Decoded Data Stream NT
170
Controller

FIG. 6

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US03/06326

A. CLASSIFICATION OF SUBJECT MATTER

IPC(7) :HO4B 1/69; HO4J 11/00
US CL :375/130; 370/208
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 375/130, 220, 284, 285; 370/208, 286, 289, 278

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

EAST

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 6,154,484 A (LEE et al) 28 November 2000 see figs 1-5 and col.2, lines 5-15 and col.4, lines 60-67 and col.7, lines 35-50 and col.14, lines 11-20 and col.22, lines 55-57	1-31
Y	US 6,141,317 A (MARCHOK et al)31 October 2000 see fig.2 and col.22, lines 44-67 and col.27, lines 1-67 and col.28, lines 1-67	1-31

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search 17 APRIL 2003	Date of mailing of the international search report 12 MAY 2003
Name and mailing address of the ISA/US Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231 Facsimile No. (703) 305-3230	Authorized officer BAYARD, EMMANUEL Telephone No. (703) 308-9573

INTERNATIONAL SEARCH REPORT	International application No. PCT/US03/06326

Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This international report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. ☐ Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. ☐ Claims Nos.:
because they relate to parts of the international application that do not comply with the prescribed requirements to such an extent that no meaningful international search can be carried out, specifically:

3. ☐ Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

1. ☐ As all required additional search fees were timely paid by the applicant, this international search report covers all searchable claims.

2. ☐ As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. ☐ As only some of the required additional search fees were timely paid by the applicant, this international search report covers only those claims for which fees were paid, specifically claims Nos.:

4. ☐ No required additional search fees were timely paid by the applicant. Consequently, this international search report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

☐ The additional search fees were accompanied by the applicant's protest.

☐ No protest accompanied the payment of additional search fees.

PCT

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International Bureau

INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

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H04M 11/00, H04B 15/00, 1/10
H04L 5/00, 25/08, H04B 1/10

A1

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(43) International Publication Date: 4 December 1986 (04.12.86)

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(22) International Filing Date: 5 May 1986 (05.05.86)

(31) Priority Application Number: 736,200

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(33) Priority Country: US

(71) Applicant: TELEBIT CORPORATION [US/US]; 10440 Bubb Road, Cupertino, CA 95014 (US).

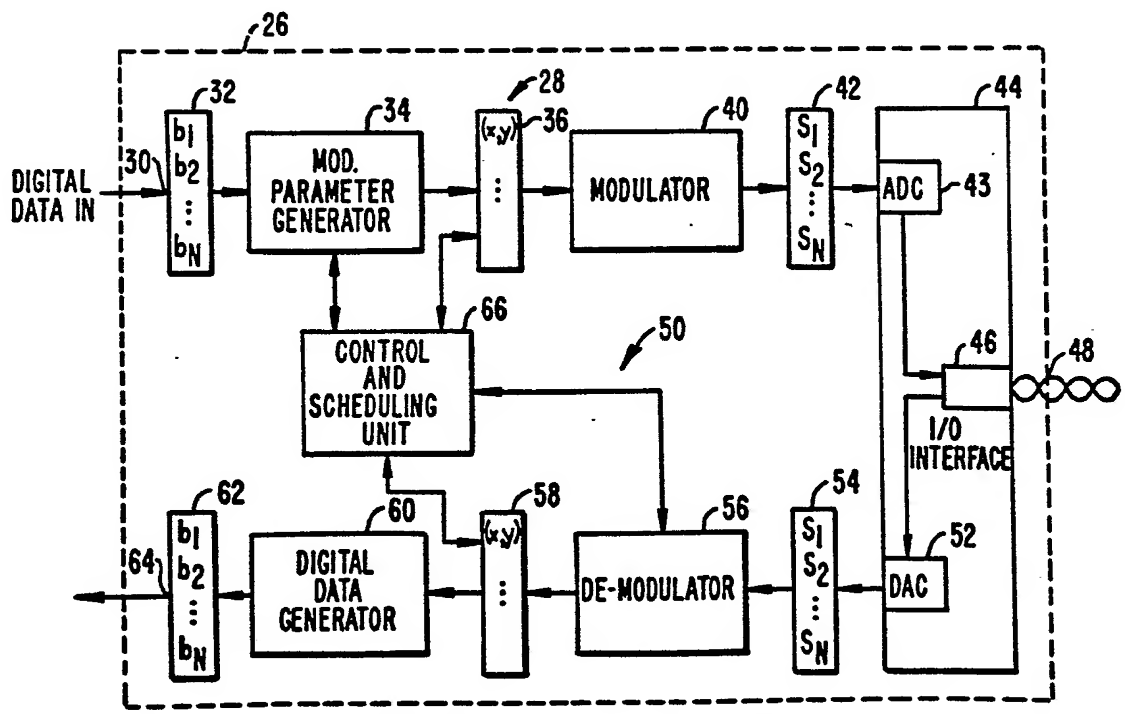
(72) Inventor: HUGHES-HARTOGS, Dirk ; 2220 Rolling Hills Drive, Morgan Hill, CA 95037 (US).

(74) Agent: ALLEN, Kenneth, R.; Townsend and Townsend, One Market Plaza, San Francisco, CA 94105 (US).

(81) Designated States: AT (European patent), AU, BE (European patent), BR, CH (European patent), DE (European patent), DK, FR (European patent), GB (European patent), IT (European patent), JP, KR, LU (European patent), NL (European patent), NO, SE (European patent).

Published
With international search report.

(54) Title: ENSEMBLE MODEM STRUCTURE FOR IMPERFECT TRANSMISSION MEDIA



(57) Abstract

A high speed modem (26) that transmits and receives digital data on an ensemble of carrier frequencies spanning the usable band of a dial-up telephone line (48). The modem includes a system (30, 32, 34, 36, 40, 43, 44) for variably allocating data and power among the carriers to compensate for equivalent noise and to maximize the data rate. Additionally, systems for eliminating the need for an equalization network, for adaptively allocating control of a channel, and for tracking variations in line parameters are disclosed.

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ENSEMBLE MODEM STRUCTURE FOR IMPERFECT TRANSMISSION MEDIA

BACKGROUND OF THE INVENTION

1. Field of the Invention:

The invention relates generally to the field of data communications and, more particularly, to a high speed modem.

2. Description of the Prior Art:

Recently, specially designed telephone lines for the direct transmission of digital data have been introduced. However, the vast majority of telephone lines are designed to carry analog voice frequency (VF) signals. Modems are utilized to modulate VF carrier signals to encode digital information on the VF carrier signals and to demodulate the signals to decode the digital information carried by the signal.

Existing VF telephone lines have several limitations that degrade the performance of modems and limit the rate at which data can be transmitted below desired error rates. These limitations include the presence of frequency dependent noise on the VF telephone lines, a frequency dependent phase delay induced by the VF telephone lines, and frequency dependent signal loss.

Generally, the usable band of a VF telephone line is from slightly above zero to about four kHz. The power spectrum of the line noise is not uniformly distributed over frequency and is generally not determinative. Thus, there is no *a priori* method for determining the distribution of the noise spectrum over the usable bandwidth of the VF line.

Additionally, a frequency-dependent propagation delay is induced by the VF telephone line. Thus, for a complex multi-frequency signal, a phase delay

between the various components of the signal will be induced by the VF telephone line. Again, this phase delay is not determinative and must be measured for an individual VF telephone line at the specific time that transmission takes place.

Further, the signal loss over the VF telephone line varies with frequency. The equivalent noise is the noise spectrum component added to the signal loss component for each carrier frequency, where both components are measured in decibels (dB).

Generally, prior art modems compensate for equivalent line noise and signal loss by gear-shifting the data rate down to achieve a satisfactory error rate. For example, in U.S. patent 4,438,511, by Baran, a high speed modem designated SM9600 Super Modem manufactured by Gandalf Data, Inc., is described. In the presence of noise impairment, the SM9600 will "gear shift" or drop back its transmitted data rate to 4800 bps or 2400 bps. The system described in the Baran patent transmits data over 64 orthogonally modulated carriers. The Baran system compensates for the frequency dependent nature of the noise on the VF line by terminating transmission on carriers having the same frequency as the frequency of large noise components on the line. Thus, Baran gracefully degrades its throughput by ceasing to transmit on carrier frequencies at the highest points of the VF line noise spectrum. The Baran system essentially makes a go/no go decision for each carrier signal, depending on the distribution of the VF line noise spectrum. This application reflects a continuation of the effort initiated by Baran.

Most prior art systems compensate for frequency dependent phase delay induced by the VF line by an equalization system. The largest phase delay is induced in frequency components near the edges of the usable band. Accordingly, the frequency components near the center of the band are delayed to allow the

frequency components at the outside of the band to catch up. Equalization generally requires additional circuitry to accomplish the above-described delays.

A further problem associated with two way transmission over the VF telephone line is that interference between the outgoing and incoming signals is possible. Generally, separation and isolation between the two signals is achieved in one of three ways:

(a) Frequency multiplexing in which different frequencies are used for the different signals. This method is common in modem-based telecommunication systems.

(b) Time multiplexing, in which different time segments are used for the different signals. This method is often used in half-duplex systems in which a transmitter relinquishes a channel only after sending all the data it has. And,

(c) Code multiplexing, in which the signals are sent using orthogonal codes.

All of the above-described systems divide the space available according to constant proportions fixed during the initial system design. These constant proportions, however, may not be suitable to actual traffic load problem presented to each modem. For example, a clerk at a PC work station connected to a remote host computer may type ten or twenty characters and receive a full screen in return. In this case, constant proportions allocating the channel equally between the send and receive modems would greatly overallocate the channel to the PC work station clerk. Accordingly, a modem that allocates channel capacity according to the needs of the actual traffic load situation would greatly increase the efficient utilization of the channel capacity.

SUMMARY OF THE INVENTION

The present invention is a high-speed modem for use with dial-up VF telephone lines. The modem utilizes a multicarrier modulation scheme and variably allocates data and power to the various carriers to maximize the overall data transmission rate. The allocation of power among the carriers is subject to the constraint that the total power allocated must not exceed a specified limit.

In a preferred embodiment, the modem further includes a variable allocation system for sharing control of a communication link between two modems (A and B) according to actual user requirements.

Another aspect of the invention is a system for compensating for frequency dependent phase delay and preventing intersymbol interference that does not require an equalization network.

According to one aspect of the invention, quadrature amplitude modulation (QAM) is utilized to encode data elements of varying complexity on each carrier. The equivalent noise component at each carrier frequency is measured over a communication link between two modems (A and B).

As is known in the art, if the bit error rate (BER) is to be maintained below a specified level, then the power required to transmit a data element of a given complexity on a given carrier frequency must be increased if the equivalent noise component at that frequency increases. Equivalently, to increase data complexity, the signal to noise ratio, S/N, must be increased.

In one embodiment of the present invention, data and power are allocated to maximize the overall data rate within external BER and total available power constraints. The power allocation system computes the marginal required power to increase the symbol rate on each carrier from n to n + 1 information units. The

system then allocates information units to the carrier that requires the least additional power to increase its symbol rate by one information unit. Because the marginal powers are dependent on the values of the equivalent noise spectrum of the particular established transmission link, the allocation of power and data is specifically tailored to compensate for noise over this particular link.

According to another aspect of the invention, a first section of the symbol on each carrier is retransmitted to form a guard-time waveform of duration $T_E + T_{PH}$ where T_E is the duration of the symbol and T_{PH} is the duration of the first section. The magnitude of T_{PH} is greater than or equal to the maximum estimated phase delay for any frequency component of the waveform. For example, if the symbol is represented by the time series, $x_0 \ldots x_{n-1}$, transmitted in time T_E; then the guardtime waveform is represented by the time series, $x_0 \ldots x_{n-1}$, $x_0 \ldots x_{m-1}$, transmitted in time $T_E + T_{PH}$. The ratio that m bears to n is equal to the ratio that T_{PH} bears to T_E.

At the receiving modem, the time of arrival, T_0, of the first frequency component of the guard-time waveform is determined. A sampling period, of duration T_E, is initiated a time $T_0 + T_{PH}$.

Accordingly, the entire symbol on each carrier frequency is sampled and intersymbol interference is eliminated.

According to a still further aspect of the invention, allocation of control to the transmission link between modems A and B is accomplished by setting limits to the number of packets that each modem may transmit during one transmission cycle. A packet of information comprises the data encoded on the ensemble of carriers comprising one waveform. Each modem is also constrained to transmit a minimum number of packets to maintain the communication link between the modems.

Thus, even if one modem has no data to transmit, the minimum packets maintain timing and other parameters are transmitted. On the other hand, if the volume of data for a modem is large, it is constrained to transmit only the maximum limited number of packets, N, before relinquishing control to the other modem.

In practice, if modem A has a small volume of data and modem B has a large volume of data, modem B will have control of the transmission link most of the time. If control is first allocated to modem A it will only transmit the minimal number, I, of packets. Thus A has control for only a short time. Control is then allocated to B which transmits N packets, where N may be very large. Control is again allocated to modem A which transmits I packets before returning control to B.

Thus, allocation of control is proportional to the ratio of I to N. If the transmission of the volume of data on modem A requires L packets, where L is between I and N, then the allocation is proportional to the ratio of L to N. Accordingly, allocation of the transmission link varies according to the actual needs of the user.

Additionally, the maximum number of packets, N, need not be the same for each modem, but may be varied to accommodate known disproportions in the data to be transmitted by A and B modems.

According to another aspect of the invention, signal loss and frequency offset are measured prior to data determination. A tracking system determines variations from the measured values and compensates for these deviations.

According to a further aspect of the invention, a system for determining a precise value of T_0 is included. This system utilizes two timing signals, at f_1 and f_2, incorporated in a waveform transmitted from modem A at time T_A. The relative phase difference

between the first and second timing signals at time T_A is zero.

The waveform is received at modem B and a rough estimate, T_{EST}, of the time of reception is obtained by detecting energy at f_1. The relative phase difference between the timing signals at time T_{EST} is utilized to obtain a precise timing reference, T_0.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 is a graph of the ensemble of carrier frequencies utilized in the present invention.

Fig. 2 is a graph of the constellation illustrating the QAM of each carrier.

Fig. 3 is a block diagram of an embodiment of the invention.

Fig. 4 is a flow chart illustrating the synchronization process of the present invention.

Fig. 5 is a series of graphs depicting the constellations for 0, 2, 4, 5, 6 bit data elements and exemplary signal to noise ratios and power levels for each constellation.

Fig. 6 is a graph illustrating the waterfilling algorithm.

Fig. 7 is a histogram illustrating the application of the waterfilling algorithm utilized in the present invention.

Fig. 8 is a graph depicting the effects of phase dependent frequency delay on frequency components in the ensemble.

Fig. 9 is a graph depicting the wave forms utilized in the present invention to prevent intersymbol interference.

Fig. 10 is a graph depicting the method of receiving the transmitted ensemble.

Fig. 11 is a schematic diagram depicting the modulation template.

Fig. 12 is a schematic diagram depicting the quadrants of one square in the modulation template.

Fig. 13 is a schematic diagram of a hardware embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

The present invention is a modem that adaptively allocates power between various carrier frequencies in a frequency ensemble to compensate for frequency dependent line noise, eliminates the need for equalization circuitry to compensate for a frequency dependent phase delay, and provides a duplex mechanism that accounts for varying channel load conditions to allocate the channel between the send and receive modems. Additional features of the invention are described below.

A brief description of the frequency ensemble and modulation scheme utilized in the present invention is first presented with respect to Figs. 1 and 2 to facilitate the understanding of the invention. A specific embodiment of the invention is then described with reference to Fig. 3. Finally, the operation of various features of the invention are described with reference to Figs. 4 through 13.

<u>Modulation and Ensemble Configuration</u>

Referring now to Fig. 1, a diagrammatic representation is shown of the transmit ensemble 10 of the present invention. The ensemble includes 512 carrier frequencies 12 equally spaced across the available 4 kHz VF band. The present invention utilizes quadrature amplitude modulation (QAM) wherein phase independent sine and cosine signals at each carrier frequency are transmitted. The digital information transmitted at a given carrier frequency is encoded by amplitude modulating the independent sine and cosine signals at that frequency.

The QAM system transmits data at an overall bit rate, R_B. However, the transmission rate on each carrier, denoted the symbol or baud rate, R_S, is only a fraction of R_B. For example, if data were allocated equally between two carriers then $R_S = R_B/2$.

In the preferred embodiment 0, 2, 4, 5 or 6 bit data elements are encoded on each carrier and the modulation of each carrier is changed every 136 msec. A theoretical maximum, R_B, assuming a 6 bit R_S for each carrier, of 22,580 bit/sec (bps) results. A typical relizable R_S, assuming 4 bit R_S over 75% of the carriers, is equal to about 11,300 bps. This extremely high R_S is achieved with a bit error rate of less than 1 error/100,000 bits transmitted.

In Fig. 1, a plurality of vertical lines 14 separates each ensemble into time increments known hereafter as "epochs." The epoch is of duration T_E where the magnitude of T_E is determined as set forth below.

The QAM system for encoding digital data onto the various carrier frequencies will now be described with reference to Fig. 2. In Fig. 2 a four bit "constellation" 20 for the nth carrier is depicted. A four bit number may assume sixteen discrete values. Each point in the constellation represents a vector (x_n, y_n) with x_n being the amplitude of the sine signal and y_n being the amplitude of the cosine signal in the above-described QAM system. The subscript n indicates the carrier being modulated. Accordingly, the four bit constellation requires four discrete y_n and four discrete x_n values. As described more fully below, increased power is required to increase the number of bits transmitted at a given carrier frequency due to the equivalent noise component at that frequency. The receive modem, in the case of four bit transmission, must be able to discriminate between four possible values of the x_n and y_n amplitude coefficients. This

ability to discriminate is dependent on the signal to noise ratio for a given carrier frequency.

In a preferred embodiment, packet technology is utilized to reduce the error rate. A packet includes the modulated epoch of carriers and error detection data. Each packet in error is retransmitted until correct. Alternatively, in systems where retransmission of data is undesirable, epochs with forward error correcting codes may be utilized.

Block Diagram

Fig. 3 is a block diagram of an embodiment of the present invention. The description that follows is of an originate modem 26 coupled to an originate end of a communication link formed over a public switched telephone line. It is understood that a communication system also includes an answer modem coupled to the answer end of the communication link. In the following discussion, parts in the answer modem corresponding to identical or similar parts in the originate modem will be designated by the reference number of the originate modem primed.

Referring now to Fig. 3, an incoming data stream is received by a send system 28 of the modem 26 at data input 30. The data is stored as a sequence of data bits in a buffer memory 32. The output of buffer memory 32 is coupled to the input of a modulation parameter generator 34. The output of the modulation parameter generator 34 is coupled to a vector table buffer memory 36 with the vector table buffer memory 36 also coupled to the input of a modulator 40. The output of the modulator 40 is coupled to a time sequence buffer 42 with the time sequence buffer 42 also coupled to the input of a digital-to-analog converter 43 included in an analog I/O interface 44. The interface 44 couples the output of the modem to the public switched telephone lines 48.

A receive system 50 includes an analog-to-digital converter (ADC) 52 coupled to the public switched telephone line 48 and included in the interface 44. The output from the ADC 52 is coupled to a receive time series buffer 54 which is also coupled to the input of a demodulator 56. The output of the demodulator 56 is coupled to a receive vector table buffer 58 which is also coupled to the input of a digital data generator 60. The digital data generator 60 has an output coupled to a receive data bit buffer 62 which is also coupled to an output terminal 64.

A control and scheduling unit 66 is coupled with the modulation parameter generator 34, the vector table buffer 36, the demodulator 56, and the receive vector table buffer 58.

An overview of the functioning of the embodiment depicted in Fig. 3 will now be presented. Prior to the transmission of data, the originate modem 26, in cooperation with the answer modem 26', measures the equivalent noise level at each carrier frequency, determines the number of bits per epoch to be transmitted on each carrier frequency, and allocates power to each carrier frequency as described more fully below.

The incoming data is received at input port 30 and formatted into a bit sequence stored in the input buffer 32.

The modulator 34 encodes a given number of bits into an (x_n,y_n) vector for each carrier frequency utilizing the QAM system described above. For example, if it were determined that four bits were to be transmitted at frequency f_n then four bits from the bit stream would be converted to one of the sixteen points in the four bit constellation of Fig. 2. Each of these constellation points corresponds to one of sixteen possible combinations of four bits. The amplitudes of the sine and cosine signals for frequency n then corresponds to the point in the constellation encoding the four bits

of the bit sequence. The (x_n, y_n) vectors are then stored in the vector buffer table 36. The modulator receives the table of (x_n, y_n) vectors for the carriers in the ensemble and generates a digitally encoded time series representing a wave form comprising the ensemble of QAM carrier frequencies.

In a preferred embodiment the modulator 40 includes a fast Fourier transform (FFT) and performs an inverse FFT operation utilizing the (x,y) vectors as the FFT coefficients. The vector table includes 1,024 independent points representing the 1,024 FFT points of the 512 frequency constellation. The inverse FFT operation generates 1,024 points in a time series representing the QAM ensemble. The 1,024 elements of this digitally encoded time series are stored in the digital time series buffer 42. The digital time sequence is converted to an analog wave form by the analog to digital converter 43 and the interface 46 conditions the signal for transmission over the public switched telephone lines 48.

Turning now to the receive system 50, the received analog waveform from the public switched telephone lines 48 is conditioned by the interface 46 and directed to the analog to digital converter 52. The analog to digital converter 52 converts the analog waveform to a digital 1,024 entry time series table which is stored in the receive time series buffer 54. The demodulator 56 converts the 1,024 entry time series table into a 512 entry (x_n, y_n) vector table stored in the receive vector table buffer 58. This conversion is accomplished by performing an FFT on the time series. Note that information regarding the number of bits encoded onto each frequency carrier has been previously stored in the demodulator and digital data generator 60 so that the (x,y) table stored in the receive vector table buffer 58 may be transformed to an output data bit sequence by the digital data generator 60. For

example, if the (x_n, y_n) vector represents a four bit sequence then this vector would be converted to a four bit sequence and stored in the receive data bit buffer 62 by the digital data generator 60. The receive data bit sequence is then directed to the output 64 as an output data stream.

A full description of the FFT techniques utilized is described in a book by Rabiner et al., entitled Theory and Applications of Digital Signal Processing, Prentice-Hall, Inc., N.J., 1975. However, the FFT modulation technique described above is not an integral part of the present invention. Alternatively, modulation could be accomplished by direct multiplication of the carrier tones as described in the above-referenced Baran patent, which is hereby incorporated by reference, at col. 10, lines 13-70, and col. 11, lines 1-30. Additionally, the demodulation system described in Baran at col. 12, lines 35-70, col. 13, lines 1-70, and col. 14, lines 1-13 could be substituted.

The control and scheduling unit 66 maintains overall supervision of the sequence of operations and controls input and output functions.

Determination of Equivalent Noise

As described above, the information content of the data element encoded on each frequency carrier and the power allocated to that frequency carrier depends on the magnitude of the channel noise component at that carrier frequency. The equivalent transmitted noise component at frequency f_n, $N(f_n)$, is the measured (received) noise power at frequency f_n multiplied by the measured signal loss at frequency f_n. The equivalent noise varies from line to line and also varies on a given line at different times. Accordingly, in the present system, N(f) is measured immediately prior to data transmission.

The steps of a synchronization technique utilized in the present system to measure N(f) and

establish a transmission link between answer and originate modems 26 and 26' are illustrated in Fig. 4. Referring now to Fig. 4, in step 1 the originate modem dials the number of the answer modem and the answer modem goes off hook. In step 2 the answer modem transmits an epoch of two frequencies at the following power levels:

(a) 1437.5 Hz. at -3 dBR; and

(b) 1687.5 Hz at -3 dBR.

The power is measured relative to a reference, R, where, in a preferred embodiment, 0dBR = -9dBm, m being a millivolt. These tones are used to determine timing and frequency offset as detailed subsequently.

The answer modem then transmits an answer comb containing all 512 frequencies at -27dBR. The originate modem receives the answer comb and performs an FFT on the comb. Since the power levels of the 512 frequencies were set at specified values, the control and scheduling unit 66 answer modem 26 compares the (x_n, y_n) values for each frequency of the received code and compares those values to a table of (x_n, y_n) values representing the power levels of the transmitted answer code. This comparison yields the signal loss at each frequency due to the transmission over the VF telephone lines.

During step 3 both the originate and answer modems 26 and 26' accumulate noise data present on the line in the absence of any transmission by either modem. Both modems then perform an FFT on the accumulated noise signals to determine the measured (received) noise spectrum component values at each carrier frequency. Several epochs of noise may be averaged to refine the measurement.

In step 4 the originate modem transmits an epoch of two frequencies followed by an originate comb of 512 frequencies with the same power levels described above for step 2. The answer modem receives the epoch and the originate comb and calculates the timing, fre-

quency offset and signal loss values at each carrier frequency as described above for the originate modem in step 2. At this point the originate modem 26 has accumulated noise and signal loss data for transmission in the answer originate direction while the answer modem has accumulated the same data relating to transmission in the originate answer direction. Each modem requires data relating to transmission loss and receive noise in both the originate-answer and answer-originate directions. Therefore, this data is exchanged between the two modems according to the remaining steps of the synchronization process.

In step 5 the originate modem generates and transmits a first phase encoded signal indicating which carrier frequencies will support two bit transmission at standard power levels in the answer-originate direction. Each component that <u>will</u> support two bits in the answer-originate direction at a standard power level is generated as a -28 dBR signal with 180° relative phase. Each component that <u>will</u> <u>not</u> support two bit transmission in the answer-originate direction at the standard power level is coded as a -28 dBR,0° relative phase signal. The answer modem receives this signal and determines which frequency carriers will support two bit transmission in the answer-originate direction.

In step 6 the answer modem generates and transmits a second phase encoded signal indicating which carrier frequencies will support two bit transmission in both the originate-answer and answer-originate directions. The generation of this signal is possible because the answer modem has accumulated noise and signal loss data in the originate-answer direction and has received the same data for the answer-originate direction in the signal generated by the originate modem in step 5. In the signal generated by the originate modem, each frequency component that <u>will</u> support two bits in <u>both</u> directions is coded with 180° relative

phase and all other components are coded with 0° relative phase.

A transmission link now exists between the two modems. In general, 300 to 400 frequency components will support two bit transmission at a standard power level, thereby establishing about a 600 bit/epoch rate between the two modems. In step 7 the originate modem sends data on the number of bits (0 to 15) and the power levels (0 to 63dB) that can be supported on each frequency in the answer-originate direction in ensemble packets formed over this existing data link. Accordingly, both the originate and answer modem now have the data relating to transmission in the answer-originate direction. The steps for calculating the number of bits and power levels that can be supported on each frequency component will be described below.

In step 8 the answer modem sends data on the number of bits and power levels that can be supported on each frequency in the originate-answer direction utilizing the existing data link. Thus, both modems are apprised of the number of bits and power levels to be supported on each frequency component in both the answer-originate and originate-answer directions.

The above description of the determination of the equivalent noise level component at each carrier frequency sets forth the required steps in a given sequence. However, the sequence of steps is not critical and many of the steps may be done simultaneously or in different order, for example, the performance of the FFT on the originate code and the accumulation of noise data may be done simultaneously. A precise timing reference is also calculated during the synchronization process. The calculation of this timing reference will be described more fully below after the description of the method for calculating the number of bits and power levels allocated to each frequency component.

It is a common VF telephone line impairment that a frequency offset, of up to 7 Hz, exists between transmitted and received signals. This offset must be corrected for the FFT to function reliably. In a preferred embodiment, this correction is achieved by performing a single sideband modulation of the quadrature tones at the offset frequency by the true and Hilbert images of received signal. Synchronization and tracking algorithms generate estimates of the frequency offset necessary.

Power and Code Complexity Allocation

The information encoded on each carrier frequency signal is decoded at the receiver channel by the demodulator 56. Channel noise distorts the transmitted signal and degrades the accuracy of the demodulation process. The transmission of a data element having a specified complexity, e.g., B_0 bits at a specified frequency, f_0, over a VF telephone line characterized by an equivalent noise level component, N_0, will now be analyzed. Generally, external system requirements determine a maximum bit error rate (BER) that can be tolerated. For the transmission of b_0 bits at noise level N_0 and frequency f_0, the signal to noise ratio must exceed E_b/N_0 where E_b is the signal power per bit to maintain the BER below a given BER, $(BER)_0$.

Fig. 5 depicts the QAM constellations for signals of various complexities B. An exemplary signal to noise ratio, E_b/N_0, for each constellation and the power required to transmit the number of bits in the constellation without exceeding $(BER)_0$ is depicted alongside each constellation graph.

A modem operates under the constraint that the total available power placed on the public switched telephone lines may not exceed a value, P_0, set by the telephone companies and government agencies. Thus, signal power may not be increased indefinitely to compensate for line noise. Accordingly, as noise

increases, the complexity of the signals transmitted must be decreased to maintain the required BER.

Most existing modems arbitrarily gear shift the signal complexity down as line noise power increases. For example, one prior art modem reduces the transmitted data rate from a maximum of 9,600 bps to steps of 7,200 bps, 4,800 bps, 2,400 bps, 1,200 bps, and so on until the bit error rate is reduced below a specified maximum. Accordingly, the signal rate is decreased in large steps to compensate for noise. In the Baran patent, the method for reducing the transmission rate takes into account the frequency dependent nature of the noise spectrum. There, each channel carries a preset number of bits at a specified power level. The noise component at each frequency is measured and a decision is made whether to transmit at each carrier frequency. Thus, in Baran, the data rate reduction scheme compensates for the actual distribution of the noise over the available bandwidth.

In the present invention, the complexity of the signal on each frequency carrier and the amount of the available power allocated to each frequency carrier is varied in response to the frequency dependence of the line noise spectrum.

The present system for assigning various code complexities and power levels to the frequency component signals in the ensemble is based on the waterfilling algorithm. The waterfilling algorithm is an information theoretic way of assigning power to a channel to maximize the flow of information across the channel. The channel is of the type characterized by an uneven noise distribution and the transmitter is subject to a power constraint. Fig. 6 provides a visualization of the waterfilling algorithm. Referring now to Fig. 6, power is measured along the vertical axis and frequency is measured along the horizontal axis. The equivalent noise spectrum is represented by the solid line 70 and

the available power is represented by the area of the cross hatched region 72. The name waterfilling comes from the analogy of the equivalent noise function to a series of valleys in a mountain filled with a volume of water representing the assigned power. The water fills the valleys and assumes a level surface. A theoretical description of the waterfilling algorithm is given in the book by Gallagher, entitled Information Theory And Reliable Communication; J. Wiley and Sons, New York, 1968, p. 387.

It must be emphasized that the waterfilling theorem relates to maximizing the theoretical capacity of a channel where the capacity is defined as the maximum of all data rates achievable using different codes, all of which are error correcting, and where the best tend to be of infinite length.

The method utilizing the present invention does not maximize the capacity of the channel. Instead, the method maximizes the amount of information transmitted utilizing the QAM ensemble described above with respect to Fig. 1 and subject to an available power restriction.

An implementation of the waterfilling concept is to allocate an increment of available power to the carrier having the lowest equivalent noise floor until the allocatd power level reaches the equivalent noise level of the second lowest carrier. This allocation requires a scan through the 512 frequencies.

Incremental power is then allocated between the lowest two carriers until the equivalent noise level of the third lowest channel is reached. This allocation level requires many scans through the frequency table and is computationally complex.

The power allocation method used in a preferred embodiment of the present invention is as follows:

(1) Calculate the system noise at the transmitter by measuring the equivalent noise at the receiver and multiplying by transmission loss. This process for measuring these quantities was described above with respect to synchronization and Fig. 4. The system noise components are calculated for each carrier frequency.

(2) For each carrier frequency, calculate the power levels required to transmit data elements of varying complexity (in the present case, 0, 2, 4, 5, 6, and 8 bits). This is accomplished by multiplying the equivalent noise by the signal to noise ratios necessary for transmission of the various data elements with a required BER, for example one error per 100,000 bits. The overall BER is the sum of the signal error rates of each modulated carrier. These signal to noise ratios are available from standard references, and are well-known in the art.

(3) From the calculated required transmission power levels, the marginal required power levels to increase data element complexity are determined. These marginal required power levels are the difference in transmission power divided by the quantitative difference in complexity of the data elements closest in complexity.

(4) For each channel generate a two column table of marginal required power levels and quantitative differences where the units are typically expressed as Watts and bits, respectively.

(5) Construct a histogram by organizing the table of step 4 according to increasing marginal power.

(6) Assign the available transmitter power sequentially over the increasing marginal powers until available power is exhausted.

The power allocation method may be better understood through a simple example. The numbers pre-

sented in the example are not intended to represent parameters encountered in an operating system.

Table 1 sets out the power requirement, P, to transmit a data element of a selected number of bits, N_1, for two carriers A and B at frequencies f_A and f_B.

TABLE 1

Carrier A

N_1	N_2-N_1	P	MP(N_1 to N_2)
0	-	0	-
2	2	4	MP(0to2)=2/bit
4	2	12	MP(2to4)=4/bit
5	1	19	MP(4to5)=7/bit
6	1	29	MP(5to6)=10/bit

Carrier B

N_1	N_2-N_1	P	MP(N_1 to N_2)
0	-	0	-
2	2	6	MP(0to2)=3/bit
4	2	18	MP(2to4)=6/bit
5	1	29	MP(4to5)=11/bit
6	1	44	MP(5to6)=15/bit

The marginal power to increase the complexity from a first number of bits, N_1, to a second number of bits, N_2, is defined by the relationship:

$$MP(N_1 \text{ to } N_2) = \frac{P_2 - P_1}{N_2 - N_1}$$

where P_2 and P_1 are the powers required to transmit data elements of complexity N_2 and N_1. N_2-N_1 is quantitative difference in the complexity of the data elements. It is understood the BER is constrained to remain below a preset limit.

The marginal powers for f_A are less than for f_B because the equivalent noise at f_B, $N(f_B)$, is greater than the equivalent noise at f_A, $N(f_A)$.

The implementation of the allocation scheme for carriers A and B will now be described. Assume that a total number of bits, N_T, are encoded on the ensemble but that no bits have been assigned to carriers A or B. For example, $N(f_A)$ and $N(f_A)$ might be greater than the powers of those carriers already carrying the data.

In this example, the system is to allocate ten remaining available power units between carriers A and B to increase the overall data element complexity by the maximum amount.

To increase N_T by two bits requires that four units of power be allocated if channel A is utilized and that six units of power be allocated in channel B is utilized. This follows because for both channels $N_1 = 0$ and $N_2 = 2$ and MP(0 to 2) = 2/bit for channel A and MP(0 to 2) = 3/bit for channel B. Therefore, the system allocates four units of power to carrier A, encodes a two bit data element on carrier A, increases the overall signal complexity from N_T to $N_T + 2$, and has six remaining available power units.

The next increase of two bits requires six power units because MP(2 to 4) = 4/bit for carrier A and MP(0 to 2) = 3/bit for channel B. Therefore, the system allocates six units of power to carrier B, encodes a two bit data element on carrier B, increases the overall signal complexity from $N_T + 2$ to $N_T + 4$ bits, and has no remaining available power units.

As is now clear, the system "shops" among the various carrier frequencies for the lowest power cost to increase the complexity of the overall ensemble data element.

The allocation system is extended to the full 512 carrier ensemble by first generating the tables of

Table 1 for each carrier during a first pass through the frequencies.

A histogram organizing the calculated marginal required power levels for all the carriers according to increasing power is then constructed. Fig. 7 is a depiction of an exemplary histogram constructed according to the present method.

In Fig. 7 the entire table of marginal powers is not displayed. Instead, the histogram is constructed having a range of 64dB with counts spaced in 0.5dB steps. The quantitative differences between the steps are utilized as counts. Although this approach results in a slight round-off error, a significant reduction in task length is achieved. The method used to construct the histogram is not critical to practicing the invention.

Each count of the histogram has an integer entry representing the number of carriers having a marginal power value equal to the power value at the count. The histogram is scanned from the lowest power level. The integer entry at each count is multiplied by the number of counts and subtracted from the available power. The scan continues until available power is exhausted.

When the scan is completed it has been determined that all marginal power values below a given level, MP(max), are acceptable for power and data allocation. Additionally, if available power is exhausted partially through marginal power level, MP(max), then k additional carriers will be allocated power equal to MP(max + 1).

The system then scans through the ensemble again to allocate power and data to the various carriers. The amount of power allocated to each carrier is the sum of marginal power values for that carrier less than or equal to MP(max). Additionally, an amount of power equal to MP(max + 1) will be allocated if the

k MP(max + 1) values have not been previously allocated.

Timing and Phase Delay Compensation

The reconstruction of (x,y) vector table by the receive system requires 1024 time samples of the received waveform. The bandwidth is about 4kHz so that Nyquist sampling rate about 8000/sec and the time sample offset between samples is 125 microseconds. The total sampling time is thus 128 msec. Similarly, the transmit FFT generates a time series having 1024 entries and the symbol time is 128 msec.

The sampling process requires a timing reference to initiate the sampling. This timing reference is established during synchronization by the following method:

During the synchronization steps defined with reference to Fig. 4, the originate modem detects energy at the 1437.5 Hz frequency component (the first timing signal) in the answer comb at time T_{EST}. This time is a rough measure of the precise time that the first timing frequency component arrives at the receiver and is generally accurate to about 2 msec.

This rough measure is refined by the following steps. The first timing signal and a second timing signal (at 1687.5 Hz) are transmitted with zero relative phase at the epoch mark.

The originate modem compares the phases of the first and second timing signals at time T_{EST}. The 250 Hz frequency difference between the first and second timing signals results in an 11° phase shift between the two signals for each 125 microsecond time sample offset. The first and second timing signals have low relative phase distortion (less than 250 microseconds) due to their location near the center of the band. Accordingly, by comparing the phases of the two timing samples and correcting T_{EST} by the number of

time sampling offsets indicated by the phase difference, a precise timing reference, T_0, can be determined.

A further difficulty relating to timing the sampling process relates to frequency dependent phase delay induced by the VF line. This phase delay typically is on the order of 2 msec, or more, for VF telephone lines. Further, this phase delay is significantly worse near the edges of the 4kHz usable band.

Fig. 8 depicts distribution of the frequency carriers of the ensemble after undergoing frequency dependent phase delay. Referring to Fig. 8, three signals 90, 92, and 94 at frequencies f_0, f_{256}, and f_{512} are depicted. Two symbols, x_i and y_i, of length T_S are transmitted at each frequency. Note that the duration of each symbol is not changed. However, the leading edge of the signals near the edge of the band 92 and 94 are delayed relative to those signals near the center of the band 94.

Additionally, for two sequentially transmitted epochs x_i and y_i the trailing section of the first symbol x_i on signals 92 and 96, near the outer edge of the band will overlap the leading edge of the second symbol y_i on the signal 94 near the center of the band. This overlap results in intersymbol interference.

If the sampling interval is framed to sample a given time interval, T_S, then complete samples of every carrier in the ensemble will not be obtained and signals from other epochs will also be sampled.

Existing systems utilize phase correction (equalization) networks to correct for phase distortion and to prevent intersymbol interference.

The present invention utilizes a unique guard-time format to eliminate the need for an equalization network. This format is illustrated in Fig. 9.

Referring now to Fig. 9, first, second, and third transmitted symbols, represented by time series x_i, y_i, and z_i, respectively, are depicted. The wave-

forms depicted in Fig. 3 are modulated on one of the carriers at frequency f. In this example a symbol time, T_S, of 128 msec. and a maximum phase delay, T_{PH}, of 8 msec are assumed. A guard-time waveform is formed by repeating the first 8 msec. of the symbol. The guard-time waveform defines an epoch of 136 msec. For example, in the first waveform 110, (X_i), the time series of the symbol, $X_0 - X_{1023}$, is first transmitted, then the first 8 msec. of the symbol, $X_0 - X_{63}$, are repeated.

The sampling of the epoch is aligned with the last 128 msec. of the guard-time waveform (relative to the beginning of the guard-time epoch defined by those frequency components which arrive first).

This detection process is illustrated in Fig. 10. In Fig. 10 first and second guard-time waveforms 110 and 112 at f_1, near the center of the band, and f_2, near the edge of the band, are depicted. The frequency component at f_1 is the component of the ensemble that arrives first at the receiver and the component at f_2 arrives last. In Fig. 10 the second waveform 112, at f_2, arrives at the receiver at $T_0 + T_{PH}$, which is 8 msec. after the time, T_0, that the first waveform 110, at f_1, arrives at the receiver. The sampling period of 128 msec. is initiated at the time $T_0 + T_{PH}$. Thus, the entire symbol on f_2, $X_0 - X_{1023}$, is sampled. The entire symbol at f_1 is also sampled because the initial 8 msec. of that symbol has been retransmitted.

Also, intersymbol interference has been eliminated. The arrival of the second symbol, (y_i), at f_1 has been delayed 8 msec. by the retransmission of the first 8 msec. of (x_i). Thus, the leading edge of the second symbol at f_1, does not overlap the trailing edge of the first symbol at f_2.

The 8 msec. guardtime reduces the usable time-bandwidth product of the system by only about 6%. This

small decrease is due to the very long duration of each symbol relative to the necessary guardtime.

Tracking

In practice, for a given carrier, the magnitudes of the (x,y) vectors extracted during the demodulation process do not fall exactly at the constellation points but are distributed over a range about each point due to noise and other factors. Accordingly, the signal is decoded utilizing a modulation template as depicted in Fig. 11.

Referring now to Fig. 11, the template is formed by a grid of squares 113 with the constellation points 114 at the centers of the squares 113.

In Fig. 11, the vector $W = (x_n, y_n)$ represents the demodulated amplitudes of the sine and cosine signals at f_n. W is in the square 113 having the constellation point (3,3) centered therein. Accordingly, W is decoded as (3,3).

The present invention includes a system for tracking to determine changes in transmission loss, frequency offset, and timing from the values determined during synchronization.

This tracking system utilizes the position of the received vectors in the squares of the demodulation template of Fig. 11. In Fig. 12, a single square is divided into four quadrants upper left, lower right, upper right, lower right, 115, 116, 117, and 118 characterized as too fast, too slow, too big, and too little, respectively. If counts in all four quadrants over time by frequency or over frequency at one time are equal or nearly equal then the system is in alignment. That is, if noise is the only impairment, then the direction of error for the decoded vector, W, should be random.

However, if transmission loss changes by even 0.1dB the number of too small counts will vary significantly from the number of too large counts. Similarly, a large difference between the number of too fast and

too slow counts indicates a phase rotation caused by a change in the offset frequency. Thus, the differences between the too fast, too slow, and too big, too small counts is an error characteristic that tracks variations in signal loss and offset frequency.

The present invention utilizes this error characteristic to adjust the signal loss and frequency offset determined during synchronization. For each frequency an adjustment of ± .1dB or ± 1.0° is made depending on the error characteristic. Other divisions of the decoding region into distinct or overlapping subregions characterized as too fast, too slow, too big, and too little are preferred in some embodiments.

Additionally, the phase of the timing signals is tracked to allow corrections of T_0.

<u>Allocation of Channel Control</u>

The present invention further includes a unique system for allocating control of an established communication link between the originate and answer modems (hereinafter designated A and B, respectively). Each waveform comprising the encoded ensemble of frequencies forms a packet of information.

Control of the transmission link is first allocated to modem A. Modem A then determines the volume of data in its input buffer and transmits between I (a minimum) and N (a previously determined maximum) packets of data as appropriate. The predetermined number N serves as a limit and the end number of transmitted packets may be significantly less than required to empty the input buffer. On the other hand, if modem A has little or no data in its input buffer it will still transmit I packets of information to maintain communication with modem B. For example, the I packets may comprise the originate or answer comb of frequencies defined above with respect to Fig. 4 and the synchronization process.

Control of the communication link is then allocated to modem B which repeats the actions of modem A. Of course, if modem B transmits the minimum number, I, of packets it is confirming to modem A the vitality of modem B.

There is no need for the limits N on the two modems to be the same, or to restrict them from being adaptable under modem control to obtain rapid character echo or other user oriented goals.

Hardware Implementation

Fig. 13 is a block diagram of a hardware embodiment of the invention. Referring now to Fig. 13, an electronic digital processor 120, an analog I/O interface 44, and a digital I/O interface 122 are coupled to a common data bus 124. The analog I/O interface 44 interfaces the public switched telephone line 48 with the common data bus 124 and the digital interface 122 interfaces digital terminal equipment 126 with the common data bus 124.

The following components are utilized in a preferred embodiment of the invention. The analog I/O interface 44 is a high performance 12 bit coder-decoder (codec) and telephone line interface. The interface has access to RAM 132 and is controlled by supervisory microprocessor 128. The codec is a single chip combination of an analog to digital converter, a digital to analog converter, and several band pass filters.

The digital I/O interface 122 is a standard RS-232 serial interface to a standard twenty-five pin RS-232 type connector or a parallel interface to a personal computer bus.

The electronic digital processor 120, includes a supervisory processor 128, a general purpose mathematical processor 130, a 32K by 16 bit shared RAM subsystem 132, and a read only memory (ROM) unit 133, coupled to an address bus 135.

The supervisory microprocessor 128 is a 68000 data processor subsystem including a 10MHz 68000 processor and the 68000 program memory. The 32K by 16 bit program memory consists of several low power, high density, ROM chips included in the ROM unit 133.

The mathematical processor 130 is a 320 digital signal microprocessor system (DSP) including a 20MHz 320 processor, the 320 program memory, and an interface to the shared RAM system. Two high speed ROM chips, included in ROM unit 133, comprise the 8192 x 16 bit program memory.

The 320 system program memory includes programs for performing the modulation table look-up, FFT, demodulation, and other operations described above. The 68000 processor handles digital data streams at the input and output, performs tasking to and supervision of the 320 signal processor and associated analog I/O, and performs self and system test as appropriate.

The invention has been explained with respect to specific embodiments. Other embodiments will now be apparent to those of ordinary skill in the art.

In particular, the ensemble of carrier frequencies need not be limited as above-described. The number of carriers may be any power of 2, e.g. 1024, or some arbitrary number. Additionally, the frequencies need not be evenly spaced over the entire VF band. Further, the QAM scheme is not critical to practicing the invention. For example, AM could be utilized although the data rate, R_B, would be reduced.

Still further, the modulation template need not be comprised of squares. Arbitrarily shaped regions surrounding the constellation points may be defined. The tracking system was described where the squares in the modulation template were divided into four quadrants. However, a given parameter may be tracked by tracking the difference in the number of

counts in arbitrary regions defined about a constellation point.

Still further, a hardware embodiment including a supervisory microprocessor and a general purpose mathematical processor has been described. However, different combinations of IC chips may be utilized. For example, a dedicated FFT chip could be utilized to perform modulation and demodulation operations.

Still further, the information units utilized in the above description were bits. However, the invention is not limited to binary system.

Accordingly, it is therefore intended that the invention can be limited except as indicated by the appended claims.

WHAT IS CLAIMED IS:

1. In a high speed modem, for transmitting data over a telephone line, of the type that encodes data elements on an ensemble of carrier frequencies, a method for allocating data and power to the carrier frequencies, said method comprising the steps of:

determining the equivalent noise component for every carrier frequency in the ensemble;

determining the marginal power requirement to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

assigning available power to the ordered marginal powers in order of increasing power;

determining the value, MP(max) at which point the available power is exhausted; and

allocating power and data to each carrier frequency where the power allocated is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units allocated is equal to the number of marginal powers for that carrier less than or equal to MP(max).

2. The invention of claim 1 where said step of ordering comprises the steps of:

providing a table of arbitrary marginal power levels; and

rounding the value of each determined marginal power level to one of the values of the table of arbitrary marginal power levels to decrease computational complexity.

3. The invention of claim 2 wherein the step of determining equivalent noise comprises the steps of:

providing an A and a B modem interconnected by a telephone line;

establishing a communication link between said A and B modems;

accumulating line noise data during a no transmission time interval at said A and B modems;

transmitting at least a first ensemble of frequency carriers from said A modem to said B modem, where the amplitude of each carrier has a predetermined value;

receiving said first ensemble at said B modem;

measuring the amplitude of each carrier received at said B modem;

comparing the measured amplitudes at said B modem with said predetermined amplitudes to determine signal loss, in dB, at each carrier frequency;

determining the value of the component, in dB, at each carrier frequency of the accumulated noise; and

adding the signal loss at each carrier frequency to the noise component at each carrier frequency to determine equivalent noise.

4. A high speed modem of the type for transmitting a signal on a VF telephone line, comprising:

means for receiving an input digital data stream and for storing said input digital data;

means for generating a modulated ensemble of carriers to encode said input digital data, where each carrier has data elements of variable complexity encoded thereon;

means for measuring the signal loss and noise loss of the VF telephone line for each carrier; and

means for varying the complexity of the data element encoded on each carrier and the amount of power allocated to each carrier to compensate for the measured signal loss and noise level.

5. A high speed modem of the type that encodes data elements on an ensemble of carriers of different frequency, said modem comprising:

a digital electronic processor;

a digital electronic memory;

bus means for coupling said processor and said memory;

means, associated with said digital electronic processor, for

determining the equivalent noise component for every carrier frequency in the ensemble;

determining the marginal power requirements to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

assigning available power to the ordered marginal powers in order of increasing power;

determining the value, MP(max) at which point the available power is exhausted; and

assigning power and data to each carrier frequency where the power assigned is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units is equal to the number of marginal powers for that carrier less than or equal to MP(max).

6. In a high speed modem, for transmitting data in the form of a QAM ensemble of carrier frequencies on a VF telephone line, of the type that measures the magnitude of a system parameter prior to

transmission, a method for tracking deviations in the magnitude of the system parameter during the receipt of data, said method comprising the steps of:

generating QAM constellations for a plurality of carrier frequencies;

constructing a demodulation template for one of said plurality of carrier frequencies comprising a plurality of first regions with one of the points of said constellation positioned within each of said first regions;

forming a set of tracking regions where each first region has a first and second tracking region disposed therein;

demodulating said ensemble of carriers to obtain the demodulation points positioned in said set of first and second tracking regions;

counting the number of points disposed in said set of first tracking regions and the number of points disposed in said set of second tracking regions;

determining the difference in the number of counts disposed in said set of first tracking regions and disposed in said tracking regions to construct an error characteristic; and

utilizing said error characteristic to adjust the magnitude of said signal parameter during the receipt of data.

7. The invention of claim 6 wherein said step of constructing a demodulation template comprises the step of:

constraining said first regions to be in the shape of squares having said constellation points centered therein.

8. The invention of claim 7 wherein said step of forming said tracking regions comprises the step of:

dividing said squares into quadrants; and

selecting said tracking regions to be symmetrically disposed quadrants.

9. In a communication system of the type including two modems (A and B) coupled by a transmission link, each modem having an input buffer for storing data to be transmitted, a method for allocating control of the transmission link between modem A and B comprising the steps of:

allocating control of the transmission link to modem A;

determining the volume of data stored in the input buffer of modem A;

determining the number, K, of packets of data required to transmit the volume of data stored in the input buffer of modem A;

transmitting L packets of data from modem A to modem B where L is equal to I_A if K is less than I_A, where L is equal to K if K is greater than or equal to I_A, and where L is equal to N_A if K is greater than N_A so that the minimum number of packets transmitted is I_A and the maximum is N_A;

allocating control of the transmission link to modem B;

determining the volume of data in the input buffer of modem B;

determining the number, J, of packets of data required to transmit the volume of data stored in the input buffer of modem B;

transmitting M packets of data from modem B to modem A where M is equal to I_B if J is less than I_B, where M is equal to J if J is greater than or equal to I_B, and where L is equal to N_B if J is greater than N_B so that the minimum number of packets transmitted is I_B and the maximum is N_B;

where allocation of control between modem A and B is dependent on the volume of data stored in the input buffers of modems A and B.

10. In a high speed modem, for transmitting data over a telephone line, of the type that encodes data elements on an ensemble of carrier frequencies, a system for allocating data and power to the carrier frequencies, said system comprising:

means for determining the equivalent noise component for every carrier frequency in the ensemble;

means for determining the marginal power requirement to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

means for ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

means for assigning available power to the ordered marginal powers in order of increasing power;

means for determining the value, MP(max) at which point the available power is exhausted; and

means allocating power and data to each carrier frequency where the power allocated is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units allocated is equal to the number of marginal powers for that carrier less than or equal to MP(max).

11. The invention of claim 10 where said means for ordering comprises:

means for providing a table of arbitrary marginal power levels; and

means for rounding the value of each determined marginal power level to one of the values of the table of arbitrary marginal power levels to decrease computational complexity.

12. The invention of claim 11 wherein an A and B modem are connected by a telephone line and the means for determining equivalent noise comprises:

means for establishing a communication link between said A and B modems;

means for accumulating line noise data during a no transmission time interval at said A and B modems;

means for transmitting a first ensemble of frequency carriers from said A modem to said B modem, where the amplitude of each carrier has a predetermined value;

means for receiving said first ensemble at said B modem;

means for measuring the amplitude of each carrier received at said B modem;

means for comparing the measured amplitudes at said B modem with said predetermined amplitudes to determine signal loss at each carrier frequency;

means for determining the value of the component, in dB, at each carrier frequency of the accumulated noise; and

means for adding the signal loss at each carrier frequency to the noise component at each carrier frequency to determine equivalent noise.

13. In a high speed modem, for transmitting data in the form of a QAM ensemble of carrier frequencies on a VF telephone line, of the type that measures the magnitude of a system parameter prior to transmission, a system for tracking deviations in the magnitude of the system parameter during the receipt of data, said system comprising:

means for generating QAM constellations for a plurality of carrier frequencies;

means for constructing a demodulation template for one of said plurality of carrier frequencies comprising a plurality of first regions with one of the

points of said constellation positioned within each of said first regions;

means for forming a set of tracking regions where each first region has a first and second tracking region disposed therein;

means for demodulating said ensemble of carriers to obtain the modulation points positioned in said set of first and second tracking regions;

means for counting the number of points disposed in said set of first tracking regions and the number of points disposed in said set of second tracking regions;

means for determining the difference in the number of counts disposed in said set of first tracking regions and disposed in said tracking regions to construct an error characteristic; and

means for utilizing said error characteristic to adjust the magnitude of said signal parameter during the receipt of data.

14. The invention of claim 13 wherein said means for constructing a demodulation template comprises:

means for constraining said first regions to be in the shape of squares having said constellation points centered therein.

15. The invention of claim 14 wherein said means for forming said tracking regions comprises:

means for dividing said squares into quadrants; and

means for selecting said tracking regions to be symmetrically disposed quadrants.

16. In a communication system of the type including two modems (A and B) coupled by a transmission link, each modem having an input buffer for storing data to be transmitted, a system for allocating control

of the transmission link between modem A and B comprising:

means for allocating control of the transmission link to modem A;

means for determining the number, K, of packets of data required to transmit the volume of data stored in the input buffer of modem A;

means for transmitting L packets of data from modem A to modem B where L is equal to I_A if K is less than I_A but less than N_A, where L is equal to K if K is greater than or equal to I_A, and where L is equal to N_A if K is greater than N_A so that the minimum number of packets transmitted is I_A and the maximum is N_A;

means for allocating control of the transmission link to modem B;

means for determining the volume of data in the input buffer of modem B;

means for determining the number, J, of packets of data required to transmit the volume of data stored in the input buffer of modem B;

means for transmitting M packets of data from modem B to modem A where M is equal to I_B if J is less than I_B, where M is equal to J if J is greater than or equal to I_B but less than N_B, and where L is equal to N_B if J is greater than N_B so that the minimum number of packets transmitted is I_B and the maximum is N_B;

where allocation of control between modem A and B is dependent on the volume of data stored in the input buffers of modems A and B.

17. In a high speed modem communication system including two modems (A and B) coupled by a transmission link, each modem having an input buffer for storing data to be transmitted, each modem for transmitting data over a telephone line and each modem of the type that encodes data elements on an ensemble of carrier frequencies, a method of operating said modems to effi-

ciently allocate power and data to the carrier frequencies, to compensate for frequency dependent phase delay, where the maximum estimated magnitude of the phase delay is T_{PH}, to prevent intersymbol interference, to allocate control of the transmission link between modem A and modem B and for initiating a sampling interval having a given time sample offset equal to the reciprocal of the sampling frequency, said method comprising:

determining the equivalent noise component for every carrier frequency in the ensemble;

determining the marginal power requirement to increase the complexity of the data element on each carrier from n information units to n + 1 information units, n being an integer between 0 and N;

ordering the marginal powers of all the carriers in the ensemble in order of increasing power;

assigning available power to the ordered marginal powers in order of increasing power;

determining the value, MP(max) at which point the available power is exhausted;

allocating power and data to each carrier frequency where the power allocated is equal to the sum of all the marginal powers less than or equal to MP(max) for that carrier and the number of data units allocated is equal to the number of marginal powers for that carrier less than or equal to MP(max);

transmitting a symbol encoded on one of said carrier frequencies where said symbol is a predetermined time duration, T_S;

retransmitting the first T_{PH} seconds of said symbol to form a transmitted waveform of duration $T_E + T_{PH}$;

allocating control of the transmission link to modem A;

determining the volume of data stored in the input buffer of modem A;

determining the number, K, of packets of data required to transmit the volume of data stored in the input buffer of modem A;

transmitting L packets of data from modem A to modem B where L is equal to I_A if K is less than I_A, where L is equal to K if K is greater than or equal to I_A, and where L is equal to N_A if K is greater than N_A so that the minimum number of packets transmitted is I_A and the maximum is N_A;

allocating control of the transmission link to modem B;

determining the volume of data in the input buffer of modem B;

determining the number,J, of packets of data required to transmit the volume of data stored in the input buffer of modem B;

transmitting M packets of data from modem B to modem A where M is equal to I_B if J is less than I_B, where M is equal to J if J is greater than or equal to I_B, and where L is equal to N_B if J is greater than N_B so that the minimum number of packets transmitted is I_B and the maximum is N_B;

where allocation of control between modem A and B is dependent on the volume of data stored in the input buffers of modems A and B;

generating an analog waveform at modem A including first and second frequency components at f_1 and f_2;

transmitting said waveform from modem A to modem B at time T_A;

adjusting the phases of said first and and second frequency components so that their relative phase difference at time T_A is equal to about 0°;

detecting energy at frequency f_1 at modem B to determine the estimated time, T_{EST}, that said waveform arrives at modem B;

determining the relative phase difference at modem B between said first and second frequency components at time T_{EST};

calculating the number of sampling time offsets, N_I, required for the relative phase of said first and second carriers to change from 0 to said relative phase difference; and

changing the magnitude of T_{EST} by N_I sampling intervals to obtain a precise timing reference, T_0.

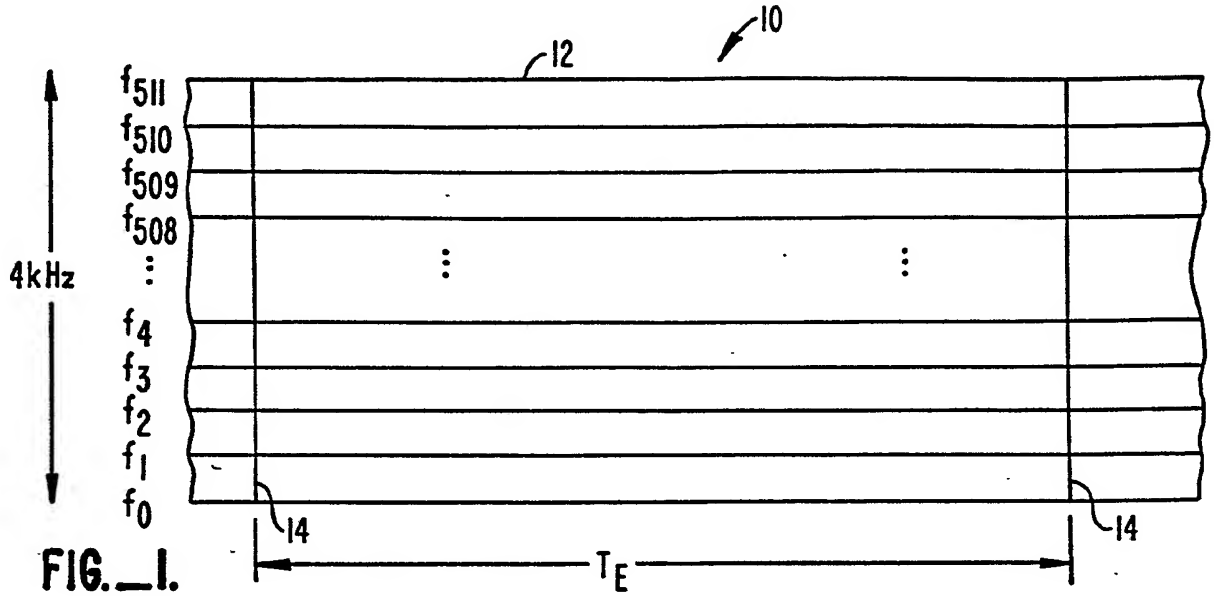

FIG._1.

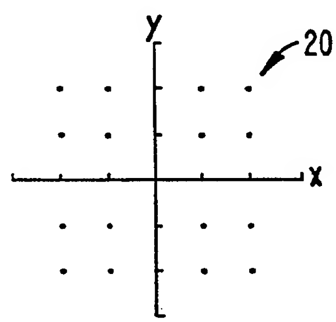

FIG._2.

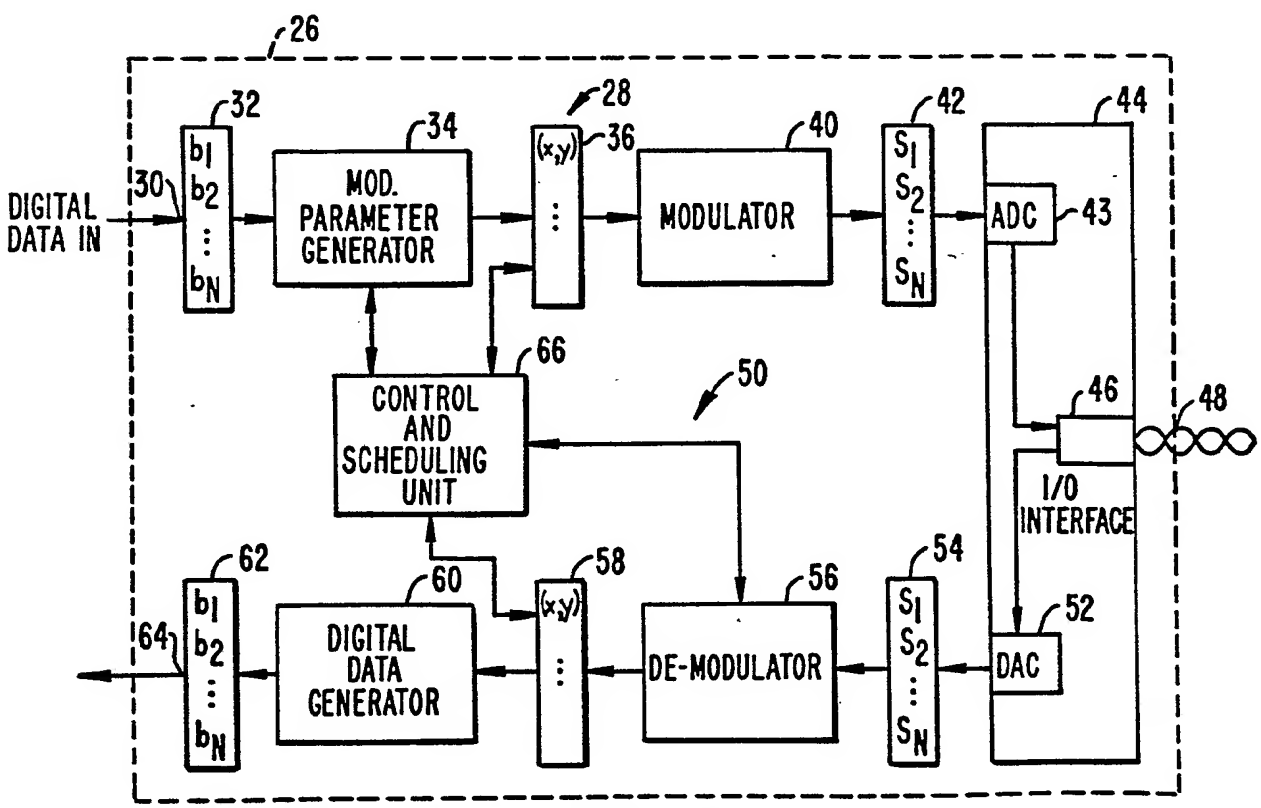

FIG._3.

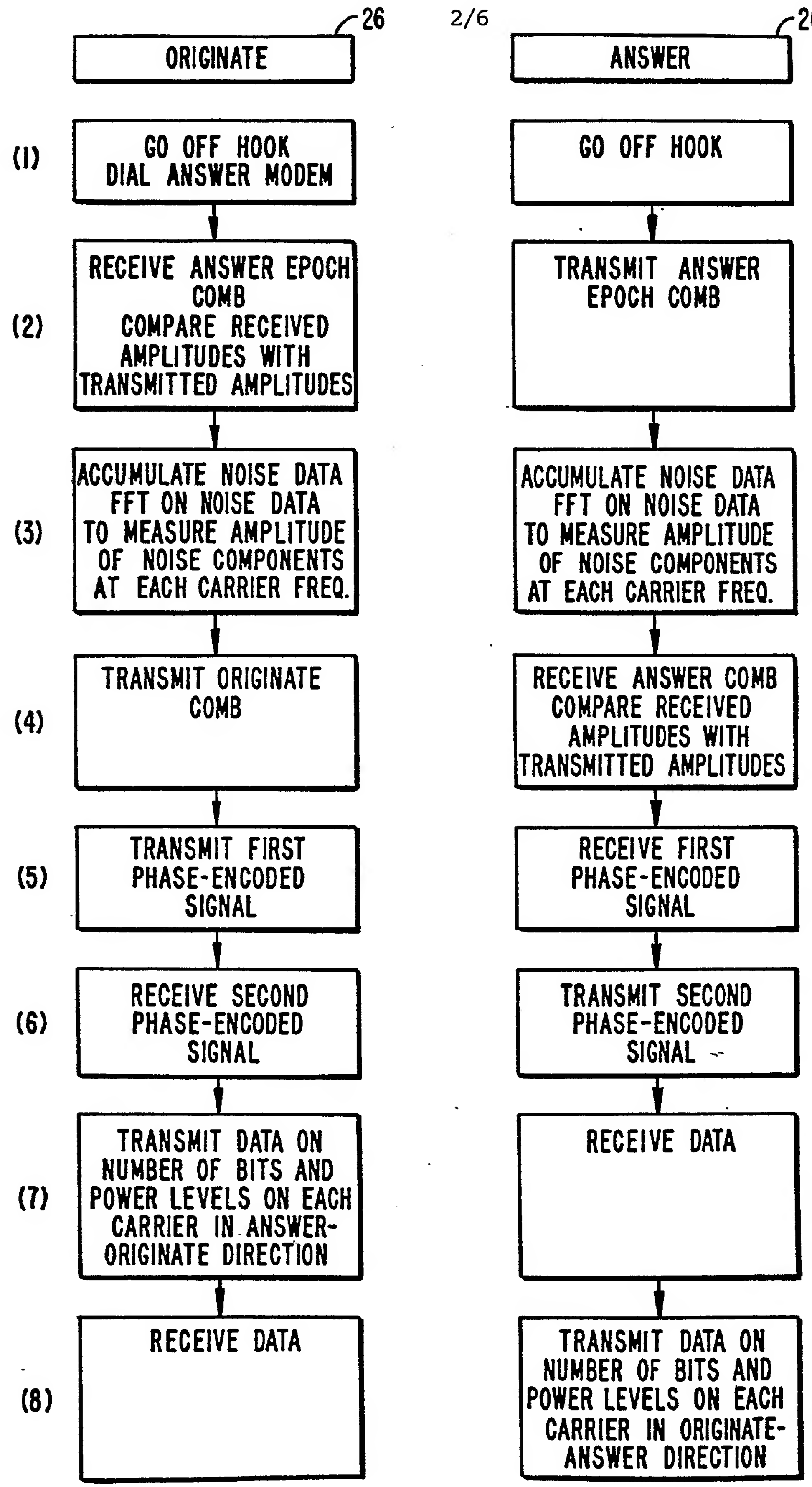
26
ORIGINATE
26'
ANSWER
(1)
GO OFF HOOK DIAL ANSWER MODEM
GO OFF HOOK
(2)
RECEIVE ANSWER EPOCH COMB COMPARE RECEIVED AMPLITUDES WITH TRANSMITTED AMPLITUDES
TRANSMIT ANSWER EPOCH COMB
(3)
ACCUMULATE NOISE DATA FFT ON NOISE DATA TO MEASURE AMPLITUDE OF NOISE COMPONENTS AT EACH CARRIER FREQ.
ACCUMULATE NOISE DATA FFT ON NOISE DATA TO MEASURE AMPLITUDE OF NOISE COMPONENTS AT EACH CARRIER FREQ.
(4)
TRANSMIT ORIGINATE COMB
RECEIVE ANSWER COMB COMPARE RECEIVED AMPLITUDES WITH TRANSMITTED AMPLITUDES
(5)
TRANSMIT FIRST PHASE-ENCODED SIGNAL
RECEIVE FIRST PHASE-ENCODED SIGNAL
(6)
RECEIVE SECOND PHASE-ENCODED SIGNAL
TRANSMIT SECOND PHASE-ENCODED SIGNAL
(7)
TRANSMIT DATA ON NUMBER OF BITS AND POWER LEVELS ON EACH CARRIER IN ANSWER-ORIGINATE DIRECTION
RECEIVE DATA
(8)
RECEIVE DATA
TRANSMIT DATA ON NUMBER OF BITS AND POWER LEVELS ON EACH CARRIER IN ORIGINATE-ANSWER DIRECTION

FIG._4.

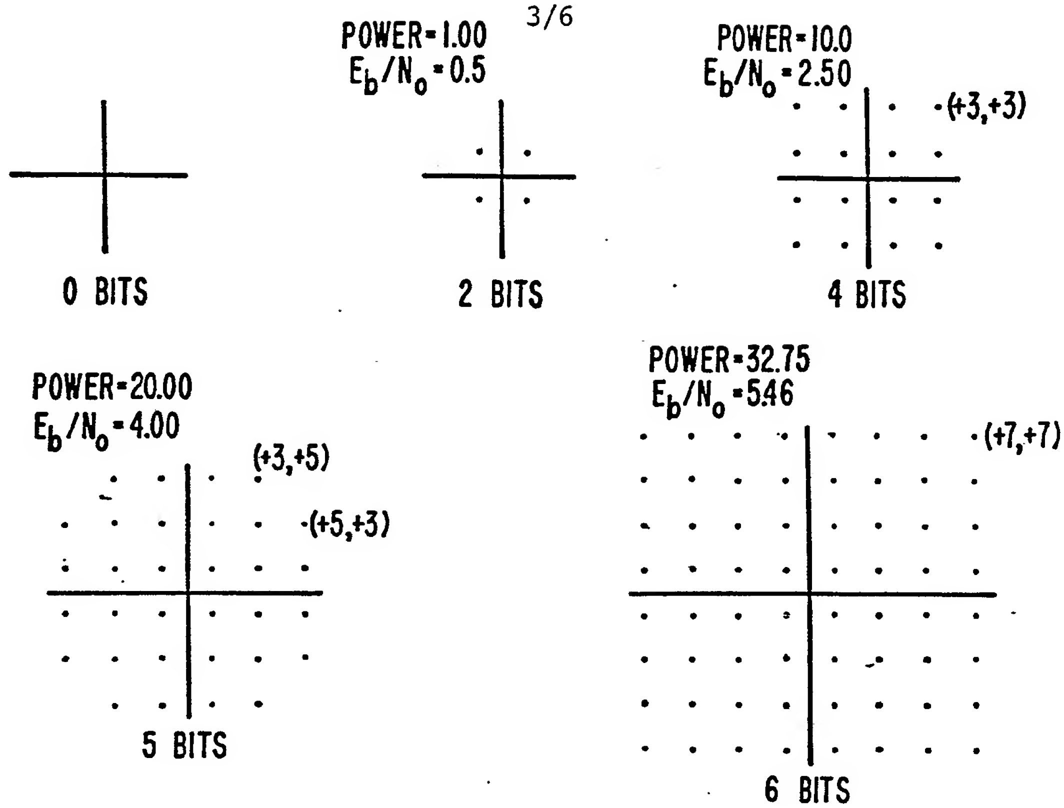
POWER=1.00
E_b/N_o=0.5
POWER=10.0
E_b/N_o=2.50
(+3,+3)
0 BITS
2 BITS
4 BITS
POWER=20.00
E_b/N_o=4.00
(+3,+5)
(+5,+3)
5 BITS
POWER=32.75
E_b/N_o=5.46
(+7,+7)
6 BITS

FIG._5.

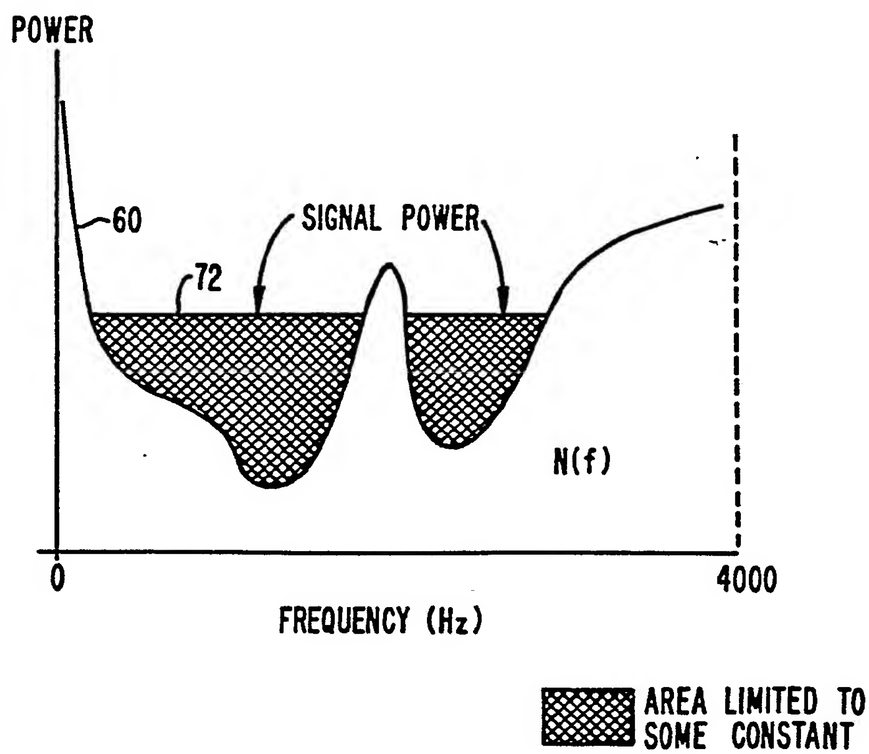
POWER
60
72
SIGNAL POWER
N(f)
0
4000
FREQUENCY (Hz)
AREA LIMITED TO SOME CONSTANT

FIG._6.

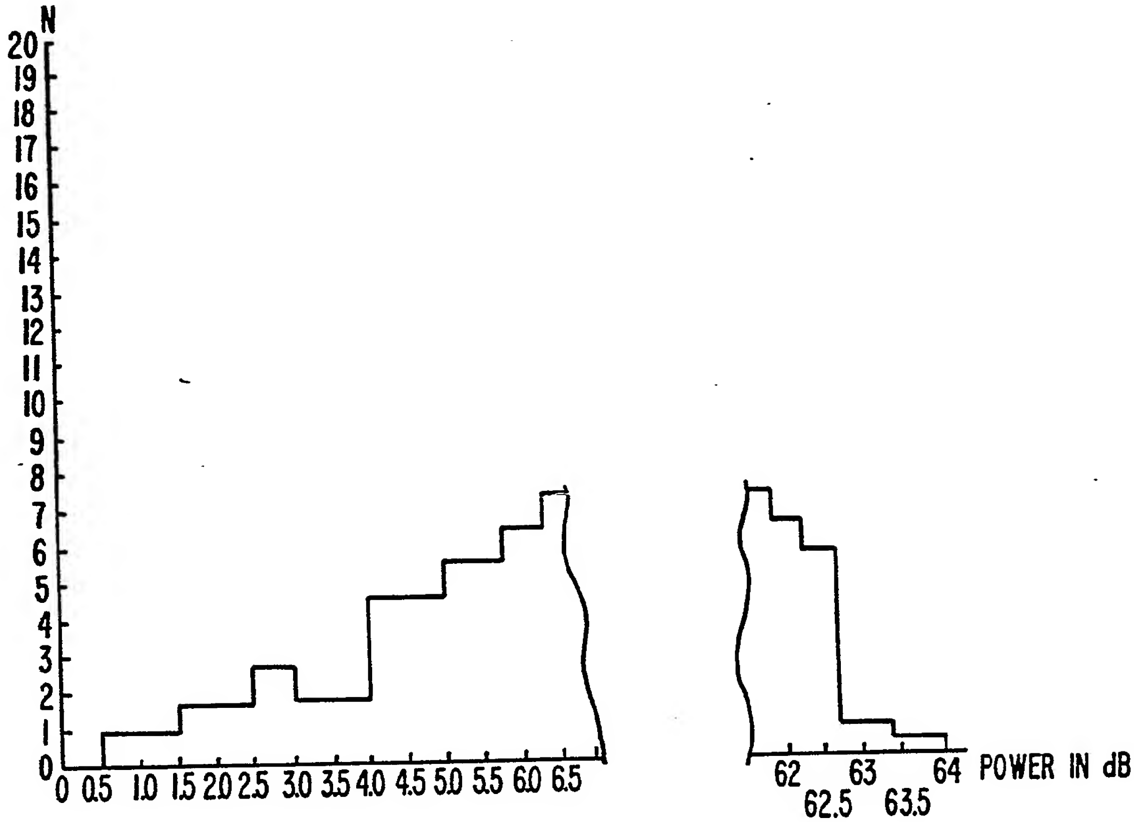
N
20
19
18
17
16
15
14
13
12
11
10
9
8
7
6
5
4
3
2
1
0
0 0.5 1.0 1.5 2.0 2.5 3.0 3.5 4.0 4.5 5.0 5.5 6.0 6.5
62 62.5 63 63.5 64
POWER IN dB

FIG.—7.

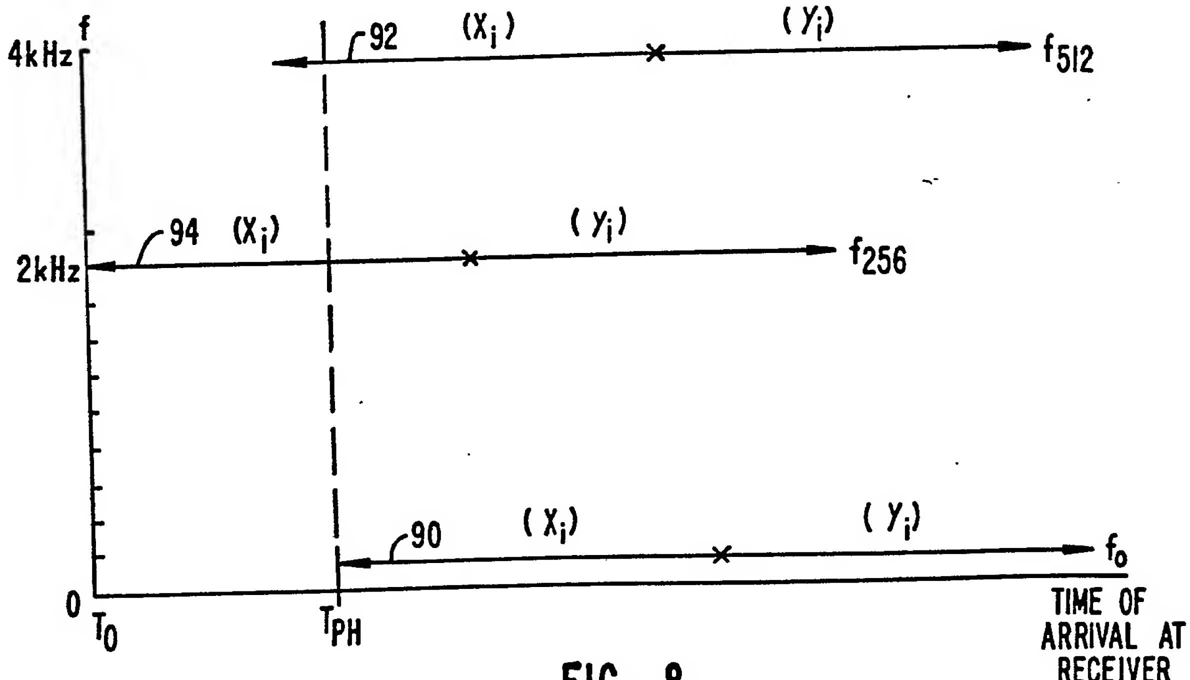
f
4kHz
2kHz
0
92
94
90
(X_i)
(Y_i)
f_512
f_256
f_o
T_0
T_PH
TIME OF ARRIVAL AT RECEIVER

FIG.—8.

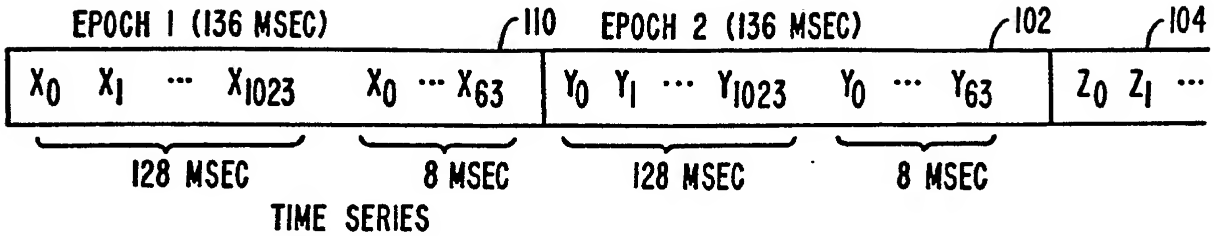

FIG._9.

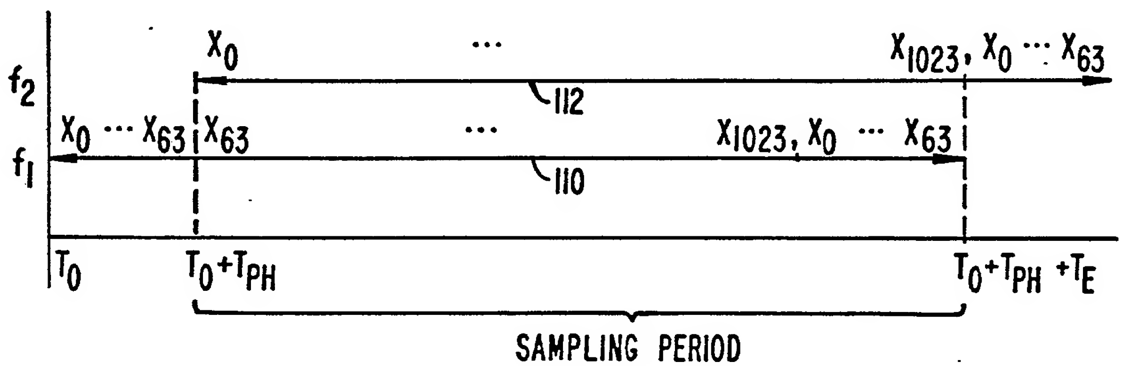

FIG._10.

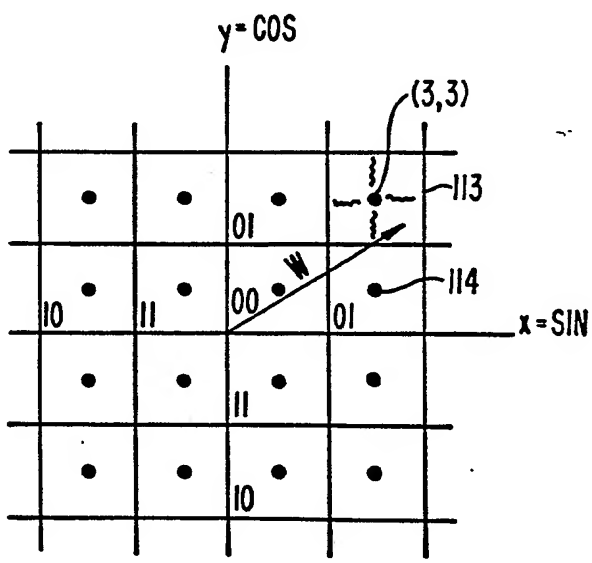

FIG._11.

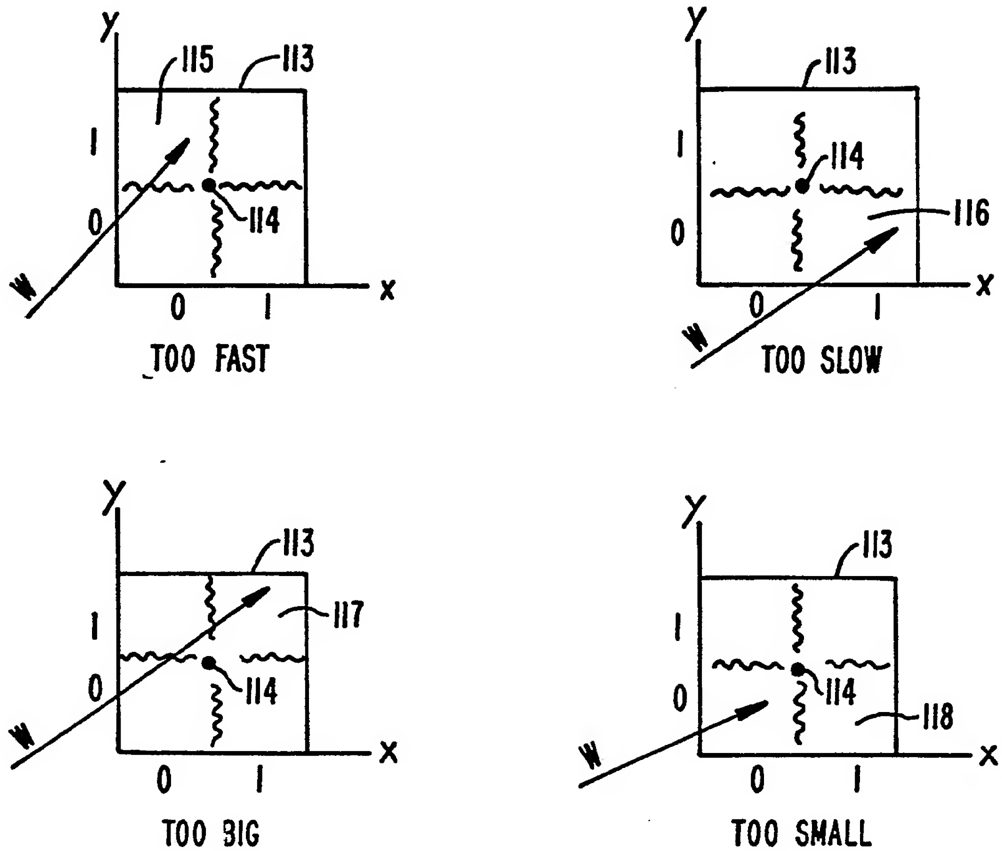
y
115
113
1
0
114
W
0
1
x
TOO FAST
y
113
114
1
0
116
W
0
1
x
TOO SLOW
y
113
1
0
117
114
W
0
1
x
TOO BIG
y
113
1
0
114
118
W
0
1
x
TOO SMALL

FIG._12.

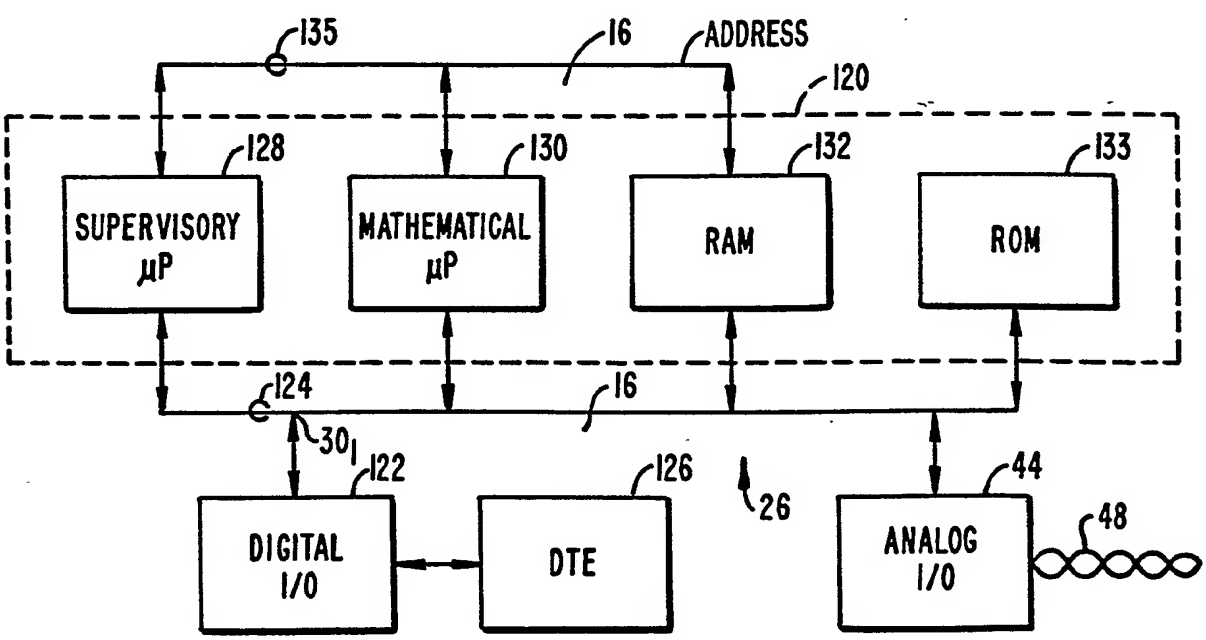
135
16
ADDRESS
120
128
130
132
133
SUPERVISORY µP
MATHEMATICAL µP
RAM
ROM
124
16
30
122
126
26
44
48
DIGITAL I/O
DTE
ANALOG I/O

FIG._13.

INTERNATIONAL SEARCH REPORT

International Application No PCT/US86/00983

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all)[3]

According to International Patent Classification (IPC) or to both National Classification and IPC

IPC (4): H04M 11/00; H04B 15/00, 1/10; H04L 5/00, 25/08; H04B 1/10
U.S. Cl.: 179/2DP; 375/39,58,99; 455/63

II. FIELDS SEARCHED

Minimum Documentation Searched[4]

Classification System	Classification Symbols
U.S.	179/2DP; 375/38,39,40,58,118; 370/16,108; 455/63,68+; 340/825.15

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched[5]

III. DOCUMENTS CONSIDERED TO BE RELEVANT[14]

Category *	Citation of Document,[16] with indication, where appropriate, of the relevant passages[17]	Relevant to Claim No.[18]
X,P	Telecommincations, Volume 19, No. 10, issued October 1985 (Dedham, Massachusetts), H.R. Johnson, "PC Communications: The Revolution Is Coming", see pages 58j to 58r.	1-17
A	US, A, 4,438,511 (Baran) 20 March 1984	1-17
A,P	US, A, 4,559,520 (Johnston) 17 December 1985	1-17
A	US, A, 4,206,320 (Keasler et al.) 03 June 1980	1-17
A	US, A, 3,810,019 (Miller) 07 May 1974	1-5,10-12,17
A	US, A, 4,328,581 (Harmon et al.) 04 May 1982	1-5,10-12,17
A	US, A, 3,971,996 (Motley et al.) 27 July 1976	6-8,13-15
A,P	US, A, 4,555,790 (Betts et al.) 26 November 1985	6-8,13-15

(cont'd)

* Special categories of cited documents:[15]

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

IV. CERTIFICATION

Date of the Actual Completion of the International Search[2]	Date of Mailing of this International Search Report[2]
17 June 1986	10 JUL 1986
International Searching Authority[1]	**Signature of Authorized Officer[20]**
ISA/US	Matthew E. Connors Matthew E. Connors

Form PCT/ISA/210 (second sheet) (October 1981)

International Application No. PCT/US86/00983

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)

Category *	Citation of Document, [16] with indication, where appropriate, of the relevant passages [17]	Relevant to Claim No [18]
A	US, A, 3,783,385 (Dunn et al.) 01 January 1974	1-5
A	US, A, 4,047,153 (Thirion) 06 September 1977	1-5
A	US, A, 4,494,238 (Groth, Jr.) 15 January 1985	1-5
A	US, A, 4,495,619 (Acampora) 22 January 1985	1-5,10-12,17
A	US, A, 4,484,336 (Catchpole et al.) 20 November 1984	1-5,10-12,17
A	US, A, 4,459,701 (Lamiral et al.) 10 July 1984	9,16,17
A	US, A, 3,755,736 (Kaneko et al.) 28 August 1973	9,16,17
A	US, A, 4,315,319 (White) 09 February 1982	1-5,10-12,17
A,P	US, A, 4,573,133 (White) 25 February 1986	1-5,10-12,17
A	US, A, 4,392,225 (Wortman) 05 July 1983	1-5,10-12,17